DESIGN AND ANALYSIS OF WIDEBAND PASSIVE MICROWAVE DEVICES USING PLANAR STRUCTURES

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Abstract

A selected volume of work consisting of 84 published journal papers is presented to demonstrate the contributions made by the author in the last seven years of his work at the University of Queensland in the area of Microwave Engineering.

The over-arching theme in the author’s works included in this volume is the engineering of novel passive microwave devices that are key components in the building of any microwave system. The author’s contribution covers innovative designs, design methods and analyses for the following key devices and associated systems:

- Wideband antennas and associated systems
- Band-notched and multiband antennas
- Directional couplers and associated systems
- Power dividers and associated systems
- Microwave filters
- Phase shifters

Much of the motivation for the work arose from the desire to contribute to the engineering of modern microwave systems that continue to evolve rapidly due to the huge interest in wideband systems for telecommunications, healthcare (diagnosis and treatment), remote sensing, medical, security and industrial imaging, wireless internet, and many other applications.

The presented work has given the author international recognition as evidenced by the large number of citations and the use of his design by several companies and many research groups around the globe. The selected works in this volume are published in the top-ranked journals in the field of microwave engineering in the world.

Also in this volume, the originality of the author’s work is proven and in the case of multi-authored papers, his role and contributions are explained.
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Chapter 1  
Introduction and Statement of Originality

1.1 Introduction

The general area of this work is Microwave Engineering

Modern microwave engineering is an exciting and dynamic field due to the explosion in demand for many new microwave-based applications that affect the daily life of nearly every person on the planet. Modern microwave engineering involves predominantly devices and circuits’ analysis and design in contrast to the electromagnetic field theory orientation years ago. Thus, the design and development of new microwave devices and systems that serve the new generations of applications is a necessity.

With the fast developments in systems that use microwave frequency bands, such as mobile telecomm, wireless medical monitoring & imaging, global positioning, satellites for data transmission, TV broadcasting, weather forecasting and remote sensing, wireless internet, healthcare (diagnosis and treatment), and even in computer engineering with bus systems working in the GHz bands, microwave engineering has been going through a period of resurgence over the last two decades. It has also undergone a radical transformation in recent years. Microwave engineering emphasis has changed from performance at all cost to minimum cost with perfect performance, and from narrowband to wideband with minimum interference from/to other systems.

Moreover, the development of solid-state microwave devices has had a dramatic impact on the microwave engineering field. The focus has shifted from non-planar bulky structures to planar configurations that are based mainly on microstrip circuits. The recent trend in using microstrip-based structures is mainly motivated by the need for mass production with very low costs using printed circuits and/or multilayer techniques.

There has been also a dramatic change in the world of ultra-wideband (UWB) microwave engineering in the recent decade. In February 2002, the Federal Communication Commission FCC, a regulatory body in the USA, issued a ruling, which was then adopted by other regulatory bodies around the world, that UWB could be used for data communications as well as for medical imaging. The band allocated to UWB is a staggering 7.5 GHz (from 3.1 GHz to 10.6 GHz), by far the largest allocation of bandwidth to any commercial terrestrial system. The UWB systems offer important advantages, such as low power consumption, high date rate, high time resolution, obstacle penetration, resistance to interference, stealthy transmission, co-existence with narrowband systems and so on. Those advantages enable a wide range of new applications in telecommunications, radar, positioning, medical and security imaging … etc.

The introduction of UWB adds significant challenges to the design of microwave devices and systems. The bandwidth of UWB sits on top of many existing allocations, such as the widely used IEEE 802.11 a/b/g standards. To avoid any harmful interference between the newly proposed UWB systems and existing systems if they co-exist nearby, the UWB antennas, for example, should meet certain strict radiation characteristics. Concerning the design of other
key passive microwave devices, such as couplers, power dividers, filters and phase shifters, the extremely wideband performance should be achieved using compact, planar and low-cost structures that meet the mass production criteria.

The applicant has been working in the area of Microwave Engineering as a consultant and design engineer with the industry and then as a researcher in the academia for more than fifteen years. The works included here represent his contributions to this field while working at the University of Queensland in the last seven years.

The works cover innovative designs and design methods for the following key passive devices and associated systems for *wideband planar microwave systems*:

- Antennas and associated systems
- Band-notched and multiband antennas
- Directional couplers and associated systems
- Power dividers and associated systems
- Microwave filters
- Phase shifters
1.2 Bibliography


1.3 Statement of Originality and Individual Contribution

The works included in this thesis have not been submitted at any other institution for any degree.

The papers set out in this submission are wholly original works and are not, to the best of the author’s knowledge, replicas of the work of others. The originality of the work is indicated by the publication of the works in journals of high repute and relevance in the field. A convenient way to establish the originality of the submitted material is by providing information on the refereeing process of the published papers.

The papers included in this volume appeared in prestigious international journals in the field of microwave engineering. These journals include IEEE Transactions on Microwave Theory and Techniques, IEEE Transactions on Antennas and Propagation, IEEE Microwave and Wireless Components Letters, IEEE Antennas and Propagation Letters, IET Microwaves, Antennas and Propagation, IET Electronics Letters, and Wiley Microwave and Optical Technology Letters. The editorial boards of those journals consist of international reviewers who are experts in their particular fields. The submitted papers were examined in a strict refereeing process with three to five reviewers and were accepted for publication on merits of originality.

In terms of the author’s initiation of, and contributions to, the works, the following general indicators are appropriate. In the case of single-authored papers, the sole contribution is obvious unless some credit to the work of others is highlighted in the body of a paper. The number of selected papers included in this volume in which the candidate is the sole author is 4.

In cases where the candidate is the first listed author, he played a large part in initiating the research, performed much of the theoretical and applied research and wrote the paper. As an average estimation of contribution in the selected papers, where the candidate is the first author, at least 60% of the total work is attributable to him. The number of the selected papers in which the candidate is the first author is 19 papers.

The number of the selected papers in this volume in which the candidate is not the first author is 21 papers. Where Prof. Bialkowski, who was the candidate’s host and co-worker when he was a Postdoctoral Research Fellow at the University of Queensland, is the first author on a paper, the research was initiated by the candidate except the research on six-port devices which was initiated by Prof. Bialkowski. The candidate designed the devices and did all the simulations and measurements, whereas Prof. Bialkowski wrote most of the paper. In the other selected papers where the candidate is not the first listed author, one of the candidate’s PhD students is the first author. In such cases, the research was initiated by the candidate who wrote most of the paper, whereas the students did the simulations and measurements. In instances where the candidate was not the first author on a paper, the amount of contribution varies from 30% up to 50% of the work.

It is difficult to place a numerical amount on the overall contribution of the author to this body of work. In light of the above comments, however, it is stipulated that he is the main contributor and played the leading role in all the works described herein. The author’s contributions thus represent more than 80% of the works included in this volume.
Chapter 2 Specific Areas of Contributions

2.1 Index to Papers by Theme

The material included in this volume is the collection of 84 selected journal papers produced by the author in the field of wideband passive microwave devices. The topics covered by those papers can be broken into the following six theme areas of Microwave Engineering. Listed under each theme are the selected journal papers published in that theme arranged in chronological order. In the subsequent sections, a description of each area is given followed by short reviews of individual papers.

**Antennas and associated systems**


**Band-notched and multiband antennas**


**Directional couplers and associated systems**


**Power dividers and associated systems**


**Microwave filters**


**Phase shifters**


2.2 Antennas & Associated Systems (Papers# 1 to 24)

The performance of any wireless system is heavily dependent on the design of its antenna which is effectively the “ear and eye” of that system. The explosive growth of broadband wireless systems and the commercial utilization of ultra-wideband (UWB) systems in many applications have raised unique design challenges for antennas serving those systems. A great deal of research efforts has been expended by the author to achieve the desired broadband characteristics in terms of impedance matching, group delay and radiation properties, as well as other practical requirements, such as low cost and miniaturized size. The requirements placed on wideband antennas in terms of size, phase linearity and spectral efficiency are more demanding than for narrowband antennas.

The author develops the theory, analysis, and design procedure for different types of planar omnidirectional and directional wideband antennas as explained in [1]-[12], [15], [23], [24]. The theories behind the design of those antennas are discussed in details. Closed-form procedures are derived to enable finding the values of the different design parameters of the antennas for a certain performance. Those proposed procedures enable the designers to avoid the trial-and-error approach that is usually combined with brutal-force simulations; a process which consumes a great deal of time and effort. The accuracy of the proposed methods is validated by developing many types of antennas that demonstrate high quality performance (gain, radiation pattern, group delay and efficiency) with compact size and low cost. The structures of some of the proposed antennas include special features for improved cut-off at the out-of-band frequencies [10] or for an extremely compact size [11], [13]. Some of the developed antennas are used in different microwave imaging systems [12], [16]-[18], [21], [22].

The author also investigates the design of several types of microstrip reflectarrays that have wideband performance [14], [19], [20]. The reflectarray is an antenna that consists of a flat reflecting surface with many microstrip elements and a feed antenna. It uses a suitable phasing scheme to convert a spherical wave produced by its feed into a plane wave. The microstrip reflectarray is a high gain antenna which evolved as an efficient and cost-effective replacement of the parabolic reflectors and phased arrays: The parabolic reflector lacks the ability to achieve wide angle beam scanning, whereas the high gain phased array with electronic scanning is very expensive due to its complicated beamforming network and amplifier modules.

A brief explanation of the contents of the papers is given hereafter in a chronological order.

In [1]-[3], the author presents novel procedures for designing directive tapered-slot and omnidirectional antennas. The antennas operate over the UWB frequency band from 3.1 GHz to more than 10.6 GHz. In [4], the effect of the tapering profile of tapered slot antennas on the gain and bandwidth is investigated. A microwave imaging system that employs the designed tapered slot antennas and aimed at breast cancer detection is explained in [5].
A coplanar waveguide fed planar antenna with an extremely broad bandwidth in excess of 128% is presented in [6]. This comfortably covers the required bandwidth for UWB communication applications, exhibits the required omnidirectional pattern characteristics and has a compact size.

Two types of coplanar waveguide fed quasi-Yagi antennas with broad bandwidth, high front-to-back ratio and high efficiency are presented in [7] and [8]. The uniqueness of the design in those antennas is the simple feeding structure. Despite that simplified structure, the antennas achieve more than 44% covering the X-band. The antennas are compatible with microstrip circuitry and active devices.

An efficient approach is described in [9] for designing UWB antennas in the form of planar monopoles of elliptical and circular shape. The presented results show that the proposed method can be applied directly to design planar antennas with UWB behavior, omnidirectional characteristics, good radiation efficiency and high fidelity factor.

The paper [10] describes a method to improve the cutoff capability of UWB planar antennas at the out-of-band frequencies using a meandered slot. In the presented design, the antenna is formed by a planar monopole and a ground plane both of half circle shape, with a meandered-shape slot made in the monopole. The results show that the meandered slot improves the cutoff capability of the antenna by increasing the return loss in the lower and upper out-of-bands by more than 5 dB without any negative effect on the passband.

A procedure to design miniaturized planar UWB omnidirectional antennas is explained in [11]. The proposed method utilizes corrugated radiator and ground plane of elliptical shapes to design antenna of compact size. The proposed method results in a surface area reduction by more than 50% compared with the optimized non-corrugated structure.

A compact and directive UWB antenna is presented in [12]. The antenna is in the form of an antipodal tapered slot with resistive layers to improve its directivity and to reduce its backward radiation. The time domain performance of the antenna shows negligible distortion which makes it suitable for imaging systems. To show the possibility of using the proposed antenna in time-domain imaging system, the use of the proposed antenna for breast imaging is also studied.

A method to design a microstrip-fed antipodal tapered-slot antenna, which has UWB performance and miniaturized dimensions, is described in [13]. The proposed structure removes the need to use any transitions and/or baluns in the feeding structure and enables the direct connection between the microstrip feeder and the radiator. To miniaturize the antenna and reduce the size by more than 80% compared with traditional design, the radiator and ground plane are corrugated.

The effect of thickness of the conductive coating on performance of microstrip reflectarrays of wideband performance is investigated in [14]. The unit cell used in work is in the form of a
cross shaped ring operating at the X-band (8 GHz-12 GHz). It is proven that increasing thickness of the conductive layer decreases the losses significantly, reduces the phase slope and shifts the resonant frequency to a higher value, while it has a negligible effect on the phase range.

A planar antenna of tapered slot configuration for use in UWB microwave imaging systems aimed for early breast cancer detection is presented in [15]. It is designed to operate in a liquid of a high dielectric constant that matches the electric properties of average breast tissues. It is designed with a corrugated radiator to have a very compact size with overall dimensions of 0.9 cm×1 cm. The antenna shows a distortion-less response in the time domain and thus it is suitable for the microwave imaging systems utilizing a short-pulse radar technique, such as the strain imaging system presented in [16].

An exponentially tapered slot antenna fed using a tapered microstrip line with a suitable microstrip-slot transition is described in [17]. The microstrip-slot transition that is needed to achieve strong coupling between the microstrip feeder and the tapered slot radiator uses virtual open and short circuits in the form of a radial slot stub and a radial microstrip stub. The antenna is used to build high resolution hemispherical scanning system for breast imaging.

The design of a compact tapered slot antenna immersed in a suitably designed coupling liquid for a microwave-based brain imaging is presented in [18]. To miniaturize the antenna, corrugations are introduced in outer edges of both the radiator and the ground plane. To protect the antenna from the adverse effects, such as corrosions of the conductive layers by the coupling liquid, the antenna is covered by a dielectric sheet. The antenna covers the band from 1 GHz to 4 GHz with moderate gain.

The design of a single-layer reflectarray antenna, which employs a novel phasing element in the form of a fixed-size circular ring and a variable-length open-circuited stub, is presented in [19]. The array is developed on a thin substrate supported by a thick foam material. An X-band offset fed 13×13 element reflectarray pointing at 20° from the broadside direction is designed using the proposed structure to realize a 17.8% bandwidth.

To offer a low cost alternative to phased arrays, an electronically controlled phasing element for a beam-steered single-layer microstrip reflectarray antenna operating at 4 GHz is presented in [20] using the phasing element proposed in [19]. The reconfigurable design is accomplished by including an open gate PIN transistor in the variable length stub to offer a beam steering ability.

A planar antenna array that includes 12 of the antenna elements described in [18] is built for breast imaging system [21], [22]. The use of the compact planar array enables using simple post-processing tools to get clear three-dimensional images of the target as depicted in [21].
The author presents a closed-form method that accurately calculates the effective permittivity as seen by printed center-fed dipoles [23]. That accurate prediction enables the correct choice of the required dimensions of the dipole for a certain resonant frequency. For quasi Yagi-Uda antennas that usually have the driven element as a center-fed dipole, this means an accurate prediction of the required length of the driver, and subsequently all the other dimensions of the antenna without the need for extensive simulations and optimizations. The results obtained by full-wave electromagnetic simulations and measurements over a wide range of design parameters prove the validity of the derived method which is based on the conformal mapping.

The author presents in [24] a quasi-Yagi antenna that has an ultra-wideband performance. The design of the antenna is based on the theory explained in [23]. To enable the wideband performance, the antenna utilizes a dual-resonant driver and a balun formed using a stepped-impedance coupled structure. The driver is designed to be dual-resonant by loading it with an inductor in the form of short section of narrow microstrip line at a certain position. The balun includes a T-junction of microstrip lines and two pairs of stepped-impedance coupled lines. The designed balun covers the band from 3 GHz to 12 GHz with $180^\circ \pm 15^\circ$ differential phase between the two balanced outputs and less than 0.5 dB insertion loss. The integrated antenna shows less than -10 dB reflection coefficient, 3.6-4.5 dBi gain, 13-17 dB front-to-back ratio, less than -19 dB cross-polarization across more than 75% fractional bandwidth centered at 7.5 GHz.

2.3 Band-Notched and Multiband Antennas (Papers# 25 to 34)

As indicated previously, the UWB technology has received a huge interest due to its many desired features and capabilities. Because of the existence of other wireless standards operating across parts of the UWB spectrum, an additional requirement for UWB antennas that are used for wireless communications is to reject the undesired bands within the ultra-wide passband and thus to cancel any possibility of interference with those other narrowband systems. To that end, UWB antennas with notched characteristics at certain bands are desired. In the papers [25]-[30], [34], the author demonstrates how to achieve that requirement by introducing the band rejection function within the UWB antenna structure rather than the traditional approach of extending the antenna’s structure to include more devices, such as bandstop filters. The proposed method does not need any additional space and elements for the antenna, and thus it does not complicate the structure or increase the cost. Instead, it modifies the available space in the feeder, radiator or ground plant to realize the required performance. The proposed design method is a technique that can be utilized for different types of frequency ranges and available printed circuit boards.

On the other hand, the rapid growth in telecommunication systems means that many functions are to be integrated into a single device, such as a mobile handset. A single handset is now required to deal with multi-standard services such as voice, data, video broadcasting,
and digital multimedia. This development has led to a great demand for compact multiband antennas designed specifically to handle multi-standard services. As a response to that demand, the author presents different configurations for the radiator and the ground plane to cover the assigned bands for the multi-standard systems with a sharp cutoff for the other frequency bands used by other systems [31]-[33]. The main features of the proposed method are the compact size, low cost and excellent performance that meets the industrial standards.

A brief explanation of the contents of the papers is given hereafter in a chronological order.

A closed-form design procedure for a compact planar antenna featuring UWB performance and simultaneous signal rejection in any sub-band within the UWB spectrum assigned for other applications is presented in [25]-[28]. The rejection capability is realized by embedding a tuning slot in the feeder [25], a slot in the radiator [26], inter-digital structure in the feeder [27], dual resonators in the feeder [28], or spurline in the feeder [29] of the antenna. The designed antennas feature omnidirectional characteristics and high radiation efficiency across the required band of operation, whereas the undesired bands are attenuated severely by more than 15 dB.

A method is described in [30] to reject certain bands within the passband of a UWB planar antenna using parasitic elements. In the presented design, the antenna is created by a planar monopole and a ground plane both of half circle shape, whereas parasitic elements are in the form of printed strips. Four examples of UWB antenna design are shown. The first design is without parasitic, while the remaining ones are with parasitics to reject a single narrow band, a wide band or three narrow bands. The results show that more than 10 dB gain drop is recorded in the suppressed bands.

A multi-resonance double cross element is used to design a dual-band reflectarray antenna for X- and K-bands with dual linear polarization is described in [31]. The proposed element has a single conductive layer structure which makes it easy to manufacture. The results presented in the work show that the mutual effect between the elements of the two bands is negligible. Hence, it is easy to achieve the phase compensation for each band separately.

A compact planar antenna for portable multistandard transceivers is presented in [32]. The proposed microstrip-fed antenna includes a symmetrical double G-shaped radiator and slotted ground plane. The antenna covers the standard bands; Personal Communications Service (PCS), Wireless Local Area Netwrok (WLAN), Bluetooth, Worldwide Interoperability for Microwave Access (WiMAX), High Performance Radio Local Area Network (HIPERLAN), IEEE 802.11a, and Digital Cellular Service (DCS). Another planar omnidirectional antenna for multiband operation is presented in [33]. The antenna includes a slotted ground plane, a T-shaped radiator, meandered and open circuit strips to cover the standard bands PCS, Universal Mobile Telecommunications System (UMTS), Wireless Broadband Internet (WiBro), WLAN, Bluetooth, WiMAX, and DCS.
A planar antenna omnidirectional with UWB performance and dual band-notched characteristics is proposed in [34]. The main features of the antenna are the compact dimensions and omnidirectional radiation across the whole band of operation. The radiator of the antenna is a slotted square patch. The ground plane is located at the bottom layer, which also includes a Q-shaped conductor-backed plane used to widen the impedance bandwidth. Dual band-notched characteristics are achieved by an inverted T-shaped strip inside the slotted radiator and a pair of mirror inverted L-shaped slots at the two sides of the radiator. The suppression of the undesired two sub-bands is achieved with more than 20 dB of attenuation without any impact on the desired passband.

2.4 Directional Couplers and Associated Systems (Papers# 35 to 52)

Directional couplers can be simply described as a reciprocal four-port passive network. They are used to couple part of a microwave signal from the main transmission line into the coupled line with the required amplitude, but with a quadrature phase with respect to the signal in the main line. They are commonly used in microwave instrumentations, amplitude, phase and frequency discriminators, balanced amplifiers, balanced mixers, feeding networks of antennas, and many other applications. In addition, they are essential for developing cost-effective measurement equipments. The recent trend in designing directional couplers for the abovementioned applications requires planar structure, compact size, low cost, low insertion loss, and stable phase performance across the required wide bandwidth.

The author develops clear design procedures for that crucial component of any wideband microwave system with high stability in the overall performance across the whole band of operation. One of the proposed procedures adopts the broadside-coupled multilayer technology to achieve the required ultra wideband performance with a compact and cost-effective product [35], [36]. The presented design is especially suitable for the modern multilayer structures, such as laminated multi-chip modules and low temperature co-fired ceramics. The other important feature of the proposed approach is the possibility of using the coupling factor as a controller to customize the frequency range, delay, and phase range. With that approach, it is possible to develop compact devices each cost a fraction with superseded performance and much less size compared with a commercial rack of interconnected devices. To optimize the performance of broadside couplers, the effects of using different shaped of coupled structures [43] and different substrates [47] are investigated. The proposed broadside couplers are used by the author to build several key microwave devices and sub-systems, such as multi-port sub-system [37], transitions [38], [39], [44], modulators [48], and crossovers [49], [51], [52].

The other design approach of wideband directional couplers is based on using parallel-coupled lines in combination with a slotted ground structure and/or lumped elements [40], [41], [45], [46]. This low-cost approach is perfectly suitable for building directional couplers and other related devices using the printed-circuit-board technology. The author also uses the direct-coupled branch-line structure to build distortion-less crossovers for high power
applications [50]. The author presents the complete theory for designing directional couplers utilizing parallel-coupled lines for wideband applications in [42].

A brief explanation of the contents of each paper is given hereafter in a chronological order.

A simple design method for a class of compact couplers, which offer coupling in the range of 3–10 dB over a UWB frequency band from 3.1 to 10.6 GHz, is presented in [35]. The proposed couplers are formed by two elliptically shaped microstrip lines, which are broadside coupled through an elliptically shaped slot. The method is based on the modern multilayer technology. The UWB coupling is accompanied by an isolation and return loss in the order of 20 dB or better. The proposed method in [35] is extended in [36] to cover wider bandwidths extending from 2.3 GHz to 12.3 GHz with more than 23 dB of isolation and return loss.

As an application for the multi-layer couplers proposed by the author, multi-port devices, which are the main sub-systems for microwave measuring equipments, are designed [37]. Two types of compact fully integrated six-port devices that operate across the UWB frequency range are explained. These devices designed as microwave vector voltmeters are suitable for the mass production as they are assembled without using wire vias or crossovers.

Clear design guidelines for a UWB aperture-coupled vertical microstrip-microstrip transition are presented in [38] and [39] using two different approaches. The design is suitable for the recent trend in the design of microwave monolithic microwave-integrated circuits through the use of low temperature co-fired ceramic circuits. The proposed transition uses broadside coupling between elliptical-shaped microstrip patches at the top and bottom layers via an elliptical-shaped slot in the mid-layer. The theoretical analyses included in the papers are used to derive the required dimensions for the best performance concerning the insertion loss and the return loss over the maximum possible bandwidth.

In a different approach that is suitable for compact uniplanar systems, the design of an edge-coupled quadrature directional coupler, which has a broadband performance and relaxed coupled-line spacing, is presented in [40]. A slotted ground plane is used underneath the coupled region in order to relax the requirement for a narrow slot between the coupled lines. The broadband coupling is accompanied by a high isolation and return loss across the band from 3 GHz to 10 GHz. To improve the isolation and return loss performances even further and to extend the band, a floating-potential ground plane conductor is used underneath the coupled region as explained in [41].

A comprehensive theory for the design of directional couplers utilizing parallel-coupled lines is presented in [42]. The conformal mapping based theory aims at avoiding the use of brutal-force optimization of commercial full-wave electromagnetic solvers that is very involving and does not give any physical insight into the effect of each design parameter on the overall operation of the coupled structure. In the derived theory, closed-form solutions are derived for the mode impedances of the main types of parallel-coupled microstrip lines.
A theoretical model is proposed in [43] to investigate the effect of the tapering shape on the performance of broadside-coupled directional couplers. The model also shows that a proper choice of the tapering shape can result in a significant improvement in the performance across a wide bandwidth.

A viable vertical microstrip-to-coplanar waveguide (CPW) transition that covers a six-octave bandwidth is proposed in [44]. The proposed transition utilizes the magnetic coupling in a pair of microstrip-to-slotline transitions derived from the microstrip/CPW structure. The presented device is designed following simple design guidelines.

Closed-form methods for designing a microstrip coupler that has a simple planar structure, tight coupling, practical dimensions, and UWB performance are presented in [45] and [46]. According to the proposed methods, the coupled microstrip structure is divided into three sections. A theoretical model based on the even-odd mode analysis of four-port networks is derived and used to find the optimum length and coupling factor for each of those sections to get an ultra-wideband performance. According to the model in [45], the central section is designed to have a tight coupling by utilizing a slotted ground plane and one lumped capacitor, whereas the two side sections have loose coupling. The proposed method in [46] assumes three lumped capacitors connected at the center of the coupled structure without the need to use slotted ground.

The effects of the thickness and permittivity of printed circuit boards on the performance of UWB microstrip-slot couplers are investigated in [47]. It is shown that the wideband operation of this type of couplers is predominantly dependent on the substrate’s thickness and to a lesser degree on the substrate’s permittivity. The degradation in performance is observed for an increased substrate thickness and is explained by the presence of higher order modes. When the substrate thickness is fixed to a small value and the cutoff frequency of the higher mode is outside the investigated band, there is an optimal permittivity, which offers the best performance.

The operation of a compact six-port network operating as a quadrature phase shift keying (QPSK) modulator with a wide operational bandwidth is presented in [48]. The modulation is accomplished using two novel types of single layer six-port networks. The designed six-port modulators offer accurate QPSK symbol modulation at a symbol rate of 400 Msymbols/s across an octave bandwidth.

The design of a wideband crossover that includes a pair of two-port and another pair of four-port microstrip-slotline transitions is presented in [49]. The utilized transitions are designed such that the resultant planar crossover has high isolation and return loss, and low insertion loss and deviation in the group delay across a wideband. The device has less than 0.5 dB insertion loss, more than 15 dB return loss and isolation across 40% fractional bandwidth.

A microwave crossover using a dual-mode microstrip patch is presented in [50]. The patch is designed to operate at two orthogonal modes. Each of those modes is used to couple a pair of face-to-face ports. The isolation between the two orthogonal modes is enhanced by using
symmetrical slits in the microstrip patch. The designed crossover has the main features of high-power handling capability and distortion-less response due to its extremely low group delay deviation. A prototype is designed to operate at the WLAN band with less than 0.06 ns deviation in the group delay, and less than 1 dB insertion loss across the whole band. In other approaches to extend the band of crossovers with enhanced isolation, two pairs of microstrip to coplanar waveguide transitions are utilized as explained in [51], whereas a crossover based on slotted microstrip patch is presented in [52].

2.5 Power Dividers and Associated Systems (Papers# 53 to 68)

Power dividers are used to divide microwave signals at an input port between two or more outputs with the required phase and amplitude. They are widely used in antenna feeds, balanced amplifiers, phase shifters, automatic signal level control, signal monitoring, and many other applications. Properly designed power dividers are required to have compact size, planar structure, low cost, low insertion loss, and stable phase across the required wide bandwidth.

The author extends his method used in the development of directional couplers to the design of different types of microwave power dividers with very high stability in the performance (phase and amplitude) across the whole wideband of operation. The design procedure adopts different approaches to realize UWB performance with compact and cost-effective products. One of the utilized methods by the author includes using broadside-coupled multilayer structures that are especially suitable for the modern multilayer structures as explained in [53], [54], [56]-[61], and [64]. In another approach, slotted ground structures in printed circuits for the low-cost printed circuit board technology are presented in [55], [62], [64], and [65]. The use of parallel-coupled lines to build wideband power dividers is presented in [63], [66], and [67]. Using those approaches, it is possible to develop compact devices each costing less than one tenth of the commercial rack of interconnected devices that needs a space that is tens of times larger. As an application of the proposed design techniques, a microwave sub-system multiport devise is built and tested [68].

A brief explanation of the contents of each paper is given hereafter in a chronological order.

A compact three-way power divider with UWB behavior is presented in [53]. The proposed divider utilizes broadside coupling via multilayer microstrip/slot transitions of elliptical shapes. The device has better than 17 dB return loss and 15 dB isolation across the band 3.1-10.6 GHz.

The design of a multilayer out-of-phase power divider with a UWB performance is presented in [54]. The device employs two dielectric substrates with a common ground plane. A transition from a parallel stripline to two microstrip lines is formed to divide the power equally with 180° phase difference from a stripline input port to two microstrip output ports. The proposed divider shows equal power division with high stability of phase, low insertion
losses, and fine isolation between the two output ports across the band from 3 to more than 11 GHz.

The design of a compact out-of-phase uniplanar power divider operating over a UWB frequency band is presented in [55]. To achieve an out-of-phase signal division over a large frequency range, a T-junction formed by a slotline and a microstrip line accompanied by wideband microstrip to slotline transitions are employed.

The work in [56] describes the design of a planar 180° hybrid with UWB performance. The device employs two substrates with a common ground plane and various microstrip-slot transitions to achieve in-phase and out-of-phase signal division. Simple design guidelines are used to find the required dimensions of the structure. The performance of the proposed device reveals a well balanced power split accompanied by an almost ideal 180° and 0° differential phase shift across the band from 3.1 to more than 11 GHz.

In the paper [57], the design of a planar out-of-phase power divider in microstrip/parallel stripline technology for UWB applications is presented. As for other multilayer devices, it employs two substrates with a common ground plane. A coupling between two microstrip lines on the top and bottom substrate layers through a slot in the ground plane, and a suitable transition from two microstrip lines to a parallel stripline are used to achieve an UWB operation.

A closed-form method to design arbitrary three-way power dividers with UWB performance and compact structure is explained in [58]. The proposed devices utilize a broadside-coupled structure, which has three asymmetric coupled layers. The design approach exploits the three fundamental modes of propagation: even–even, odd–odd, and odd–even, and the conformal mapping technique to find the coupling factors between the different layers. The method is used to design three-way power dividers that have different power ratios.

A broadband inphase power divider utilizing broadside microstrip-slot for the C-band (4–8 GHz) that is widely used by satellite systems is presented in [59]. The device is especially suitable for the modern multilayer technology, where the output ports are located at different layers.

The paper [60] includes the design of UWB multilayer inphase power divider with compact structure. The divider utilizes broadside coupling via a multilayer microstrip-slot configuration. The device has about 20 dB return loss, and 10 dB isolation across the frequency band 3.1–10.6 GHz. The performance of the device proposed in [60] is then improved concerning the isolation and phase imbalance by including a suitable isolation resistor as explained in [61].

A UWB uniplanar inphase power divider is presented in [62]. The proposed device utilizes a T-microstrip junction combined with an electromagnetic coupling between a slotted ground plane and an elliptical patch at the centre of the T-junction. A resistor is also used to enhance the isolation between the output ports of the power divider.
A UWB compact equal-power three-way divider is presented in [63]. The proposed device utilizes simple, three parallel-coupled microstrip lines. In order to enable the use of practical gaps between the tightly coupled lines, slotted ground plane and lumped capacitors, which are connected symmetrically between the two sidelines and the centerline, are utilized. The conformal mapping technique is employed to find the dimensions of the device. The performance of the device concerning the output power, return loss, and isolation show that it operates across the frequency band from 4 GHz to 11 GHz.

A UWB compact three-way power is presented in [64]. The device utilizes the low-cost broadside-coupled microstrip-coplanar waveguide structure. The conformal mapping technique is used to find the dimensions of the device.

The work included in [65] reports the design of a UWB quadrature power divider in uniplanar microstrip technology. The compact size and good performance across wideband make the device suitable for use in wideband balanced amplifiers. The proposed device uses the conventional Wilkinson power divider with one of its output arms equipped with a double wireless via acting as a phase adjusting circuit.

A UWB unequal-split Wilkinson power divider covering the band from 2 GHz to 12 GHz with a 2:1 split ratio is presented in [66]. To achieve the UWB characteristics, the conventional quarter-wave arms of the divider are replaced by tapered lines that are equipped with a carefully designed isolation circuit. Moreover, two extra tapered transformers are incorporated at the output ports for matching purposes as the designed divider is of unequal-split type.

The paper [67] includes the design of a tunable wideband three-way power divider that is useful in adaptive transmitting arrays. The design is based on using three stepped-impedance coupled microstrip lines that have controllable coupling factors, and thus a variable signal ratio at the three output ports. The variation in the coupling factors is achieved by using two varactor diodes that are connected between the central coupled line and each of the side lines. The biasing voltages of the varactor diodes are used to control their capacitors and thus to achieve the required output signal ratios. A signal flow analysis is used to predict the performance of the proposed device, whereas the conformal mapping technique is used to obtain the required dimensions of the coupled structure and the varactors’ capacitors.

The design of a wideband six-port network constituted by in-phase and quadrature power dividers is presented in [68]. To achieve a wideband operation of the quadrature divider, the device uses a 90° phase shifter in the form of a double vertical wireless interconnect that utilizes microstrip to coplanar waveguide transitions. The designed six-port network offers wideband performance in terms of amplitude and phase characteristics across the frequency band from 3.5 to 9 GHz.
2.6 Microwave Filters (Papers# 69 to 77)

Filters are used literally in any electrical system to provide a safe passage for the desired signals and to block the undesired signals, such as noise, interference, spurious responses, harmonics...etc. To meet the needs of modern microwave systems, microwave filters should have planar easy-to-manufacture structure, compact size, low cost, low insertion loss across their passband, and sharp cutoff at their stopbands.

Regarding wide bandpass filters (BPF), the author utilizes multilayer broadside coupled microstrip patches to achieve an ultra-wideband performance with low cost, low insertion loss, high efficiency and compact size as explained in [69], and [72]. The author’s design approach responds to the modern trend in manufacturing where the multilayer technique is much preferred from the integration, reproducibility, compactness and efficiency perspectives. Since some UWB systems require notching certain sub-bands to avoid any interference with other nearby narrowband systems, the author modified the structure of [69] to build bandstop filters [70].

In another approach that is suitable for low-cost printed-circuit board based systems, the author uses tapered resonators [71], slotted ground structures [73], and parallel-coupled lines [75] to build BPFs with UWB passband. For microwave front-ends that need lowpass filters (LPF) to cancel any undesired harmonics due to the mixers and other stages, the author designs LPFs using slotted ground structures [76].

Due to the huge interest in using multi-standard devices, there is an urgent need for planar tunable filters that can be adjusted for the required frequency of each application or standard. Thus, the author modifies the structure used in [76] to build tunable filters using a simple printed circuit technology with the latest generation of microwave switching diodes [77]. The result of the proposed approach is low cost and compact filters with the capability to cover a wide range of frequency bands.

The recent trend in designing noise-immune microwave systems requires the use of balanced differential feeders at the input and output ports of the devices. Thus, the author uses multilayer coupler structure to build balanced bandpass filter with UWB performance as explained in [74].

A brief explanation of the contents of each paper is given hereafter in a chronological order.

The design of planar BPFs with UWB behavior is presented in [69]. The proposed filters utilize broadside coupling between elliptical-shaped microstrip patches at the top and bottom layer of the filter’s structure via an elliptical slot located at the mid layer, which contains the ground plane. A theoretical model is presented to explain performance of the suggested filters. Results of calculation show that the utilized structure can be used to build UWB BPFs with a flat group delay, which makes the presented configuration a good candidate for very narrow pulse transmission/reception. A design procedure for multisection broadside-coupled filters is explained. The developed devices have a 3 dB insertion loss bandwidth from 3.1 to
10.6 GHz with less than 1 dB insertion loss at the center of the passband, a sharp cutoff stopband, and a flat group delay within the passband.

A multilayer broadside-coupled microstrip-slot-microstrip structure is used in [70] to design a bandstop filter with a wide passband for UWB applications. The design procedure for the filter is based on the conformal mapping technique and the even- and odd-mode analysis. The theoretical analysis indicates that value of the coupling factor between the top and bottom layers of the structure can be used to control the width of the stopband, whereas centre of that band can be controlled by the length of the coupled structure. To limit the passband of the proposed bandstop filter to 3.1-10.6 GHz, which is the specified bandwidth for UWB systems, a broadside-coupled bandpass filter is integrated with the device.

A compact planar BPF is designed in [71] using a tapered slot resonator that is formed using two tapered slot antennas connected in a face-to-face configuration. The tapering profile and dimensions of the resonator control the characteristics of the filter. Passbands of 30–50% centred at about 6 GHz with an insertion loss of less than 1 dB and a peak-to-peak group delay of less than 0.5 ns are achieved in two filters having different tapering profiles. The use of an open-ended series stub within the coplanar waveguide feeder of the filters extends their high stopband beyond 15 GHz.

A BPF that uses broadside-coupled structures and covers the UWB frequency range is presented in [72]. It utilizes a broadside-coupled microstrip-slot-microstrip structure with embedded low-pass filter to achieve the required UWB performance. The filter has about 0.4 dB insertion loss, more than 17 dB return loss, and less than 0.3 ns peak-to-peak deviation in the group delay across the UWB passband. The filter has a wide high cutoff band that extends beyond 20 GHz.

BPFs that cover the frequency range (3.1–10.6 GHz) are presented in [73]. The filters utilize broadside-coupled microstrip–coplanar waveguide making them suitable for printed circuit board technology. To achieve a wide upper stopband, radial slots and stepped impedance resonators are employed either to suppress or to relocate the harmonic responses outside the band of interest. The presented design procedure relies on the quasi-static analysis and conformal mapping. The proposed filters have a wide upper stopband that extends above 20 GHz, a compact size and less than 0.1 ns peak-to-peak variation in the group delay across the passband.

A broadside-coupled structure is used to design balanced BPF with ultrawideband performance as described in [74]. The top and bottom layers of the structure contain tapered microstrip patches. Those patches are coupled via tapered slots in the ground plane, which is located at the middle layer. The employed structure operates as a BPF in the differential-mode, whereas it operates as an all-stop filter in the common-mode. The filter has a passband extending across 123% fractional bandwidth, a sharp and wide upper stopband that extends beyond 20 GHz, and a sharp lower stopband. The filter suppresses the common-mode signals...
by more than 24 dB across the whole abovementioned band. The designed filter reveals a distortionless performance in the time domain with only 0.1 ns peak-to-peak variation in the group delay.

A closed-form method based on parallel-coupled structures is presented in [75] to design microstrip bandpass filters with UWB performance, wide stopband, and practical dimensions. According to the proposed method, three subsections of different lengths and coupling factors are connected to form a stepped-impedance structure. A theoretical model is derived and used to find the optimum length and coupling factor for each of those subsections for an UWB passband and suppressed second and third harmonic responses in the stopband. The required performance is realized by generating and proper positioning of three transmission zeros in the upper stopband and three transmission poles in the passband. The derived model shows that the total length of the three-subsection coupled structure is one-third of the effective wavelength at the center of the passband. The theoretical model is used to find the required design values for the whole structure. The presented method is validated by building a BPF that has a UWB passband with less than 1 dB insertion loss and a wide upper stopband that extends up to 28 GHz.

The paper [76] describes a lowpass filter utilizing a single-substrate microstrip/coplanar waveguide broadside-coupled structure. The main features of the proposed filter are the extremely wide stopband, sharp cutoff response, compact size and closed-form design procedure. The theory of operation for the proposed filter is presented, its design procedure is derived and its performance is explained. The filter shows a negligible radiation, flat group delay and a sharp and wide stopband that is larger than 14 times the 3 dB cutoff frequency of the filter. The length of the coupled structure required to build the filter is less than 6% of the effective wavelength calculated at the cutoff frequency.

The design presented in [76] is modified to make the filter tunable as explained in [77]. The tuning capability of the filter is enabled by using two varactor diodes connected between the coupled lines to change their \( \pi \)-mode impedance. The performance of a prototype shows a cutoff frequency tuning range from 1.5 GHz to 2.5 GHz and a stopband that extends to 25 GHz.

2.7 Phase Shifters (Papers# 78 to 84)

Phase shifters are common microwave devices widely used to control the phase of microwave signals in mobile satellite systems, microwave instrumentation and measurement systems, modulators, noise cancellation systems, frequency converters, electronic beam-scanning phased arrays, microwave imaging and many other industrial applications.

Phase shifters are required to have compact size, low cost, and low insertion loss across the required bandwidth. The size of the phase shifters has become a crucial parameter in their design, especially since the recent adoption of phased array in the design of portable
microwave devices, such as mobile handsets, due to the limited available space. Moreover, the cost of the utilized phase shifters should be as low as possible for obvious economical reasons. In addition, the level of the insertion loss caused by the utilized phase shifters is a key factor that defines the overall performance of modern microwave systems with large dynamic ranges. A significantly high insertion loss of phase shifters used in a transmitter causes a significant reduction in the level of transmitted power, whereas it causes a serious degradation in the signal to noise ratio when the phase shifter is part of a receiver. Both of those effects reduce the dynamic range significantly of even the best designed systems.

The author derives a closed-form method to exploit the broadside coupling between different layers of a multi-layer coupled structure to build compact and cost-effective phase shifters with low insertion loss and extremely wideband performance [78], [81]. The author develops a complete design procedure for that crucial component of any UWB system with very high phase stability in performance across the whole band of operation. To serve the need of wideband microwave systems that are based on the printed circuit technology, the author proposes different configurations using parallel coupled structures [79], or slotted ground planes [80], [82].

Due to the recent trend in designing multi-standard microwave systems, tunable phase shifters have become a crucial building element in those systems. The author presents two different configurations of tunable phase shifters that have wide phase tenability across a wide frequency band. The presented devices are based on using a combination of parallel-coupled structures and slotted ground planes as explained in the papers [83] and [84].

A brief explanation of the contents of each paper is given hereafter in a chronological order.

A method with clear guidelines is presented in [78] to design compact planar phase shifters with UWB characteristics. The proposed method exploits broadside coupling between top and bottom elliptical microstrip patches via an elliptical slot located in the mid layer, which forms the ground plane. A theoretical model is used to analyze performance of the proposed devices. The derived method is used to design 30° and 45° phase shifters that have compact size. The designed phase shifters achieve better than ±3° differential phase stability, less than 1 dB insertion loss, and better than 10 dB return loss across the UWB.

The design of a fixed phase shifter for C-band applications is presented in [79]. The proposed device is composed of two parts: a quadrature 3 dB directional coupler and the Wilkinson combiner. The quadrature coupler used in the proposed phase shifter is based on the edge-coupled microstrip lines with a slotted ground plane to make the gap required between the coupled lines feasible. A prototype device shows 45°± 5° fixed phase shift across the band 4–8 GHz with less than 1 dB insertion loss and better than 12 dB return loss at the centre of the C-band. The designed device also shows a flat group delay, which enables its use in systems with narrow pulse transmission/reception and high data rates.
A method to design planar and compact phase shifters with broadband characteristics is presented in [80] and [81]. The device included in [80] is based on the slotted ground structure, whereas the device explained in [81] uses a broadside-coupled microstrip-coplanar waveguide. A design procedure based on the conformal mapping theory is used to predict the physical dimensions of the structures for a certain phase shift. The method is used to design 45°, 60° and 90° phase shifters using the two different techniques. The developed phase shifters achieve 3 to 11 GHz bandwidth with low phase instability (±2° in one approach and ±3 in the other), very low insertion loss, high return loss, and a compact size.

The use of double microstrip-slot transitions to build a planar phase shifter is explained in [82]. The device exhibits broadband performance and offers compatibility with ordinary microstrip circuits. The method is used to develop ±90° phase shifters with more than 14 dB return loss across the band from 3.1 GHz to 11 GHz.

A complete design method for a tunable phase shifter that employs a short section of parallel-coupled microstrip lines is presented in [83]. The variation in the phase is achieved by changing the odd-mode impedance of parallel-coupled microstrip lines using a varactor diode that is connected between them. A derived theoretical model shows that a unit-cell phase shifter of around one-tenth of the guided wavelength can be utilized to achieve a continuously tunable phase range in excess of 90° depending on the required bandwidth and acceptable insertion loss. The proposed method shows that in order to achieve the required tunable phase range across a wideband, high even-mode impedance is needed. A slotted ground structure is utilized underneath the coupled structure to realize that target. The proposed method is validated by building a phase shifter that has a very small length of around one-twentieth of the wavelength with 45° tunable phase range and about 1 dB insertion loss across the band 2 GHz-2.5 GHz.

In the design of traditional reflection-type phase shifters, the coupler which represents the shifter’s backbone is usually assumed to be a quarter-wavelength 3 dB coupler. In the paper [84], the author presents a theoretical model to show that for certain values for mode impedances, a coupled structure with a length that is less than one tenth of a wavelength is enough to build a high performance reflection phase shifter. The presented analysis indicates that reflection phase shifters can be designed with a more compact size and larger phase range compared with the conventional method of using a quarter-wavelength 3 dB coupler. To realize the required mode impedances when using parallel-coupled lines, a slotted ground and one shunt chip capacitor are used. The performance of a prototype shows the capability to achieve 360° tunable phase range with less than 1.5 dB of insertion loss across 36% fractional bandwidth by using two coupled sections of total length that is less than one seventh of the wavelength.
2.8 Concluding Remarks

The last decade has seen remarkable developments in Microwave Engineering. The interest from both the industry and the academia has been increasing quite rapidly in the design of wideband passive microwave devices of planar structures in specific. That interest is underpinned by the increased need for wideband microwave systems to meet the huge range of applications that grow by the day. The works described herein represent considerable original contributions of international significance to the field. There is no doubt that the pace of development in the wideband passive microwave devices will continue into the foreseeable future.

Amin Abbosh
October, 2012.
Chapter 3
Full Text of the Selected Papers
4. CONCLUSION

A broadband and high-gain microstrip slot antenna backed by a ground plane has been presented. By choosing the design parameters of the proposed structure, the antenna achieves about 102% impedance bandwidths of VSWR ≤2 and the basically stable radiation patterns across the whole bands. Furthermore, the proposed antenna obtains high gain of −5.5–7.8 dB, which has an average enhancement of 2.5 dB compared with the slot antenna not backed by a ground plane. As this antenna has broad bandwidth, high gain, low profile, and is lightweight, it is suitable for applications in military electronic countermeasure, commercial communications, and other wideband systems.

REFERENCES


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COMPACT ULTRA-WIDEBAND PLANAR TAPERED SLOT ANTENNA FOR USE IN A MICROWAVE IMAGING SYSTEM

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ABSTRACT: The design of an ultra-wideband planar tapered slot antenna for use in a circular cylindrical microwave imaging system is pre-
sent. The antenna was designed assuming high dielectric substrate material Rogers RT6010LM to achieve its compact size. The developed antenna element (50 × 50 mm²) features a 10-dB return loss bandwidth from 2.75 GHz to more than 11 GHz. The gain of the antenna is between 3.5 and 9.4 dBi over the 3–10 GHz band. The experimental tests showed that the manufactured antenna element supports transmission of narrow pulses with negligible distortions, as required in the microwave imaging system. © 2006 Wiley Periodicals, Inc. Microwave Opt Technol Lett 48: 2212–2216, 2006; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21906

Key words: UWB antennas; microwave imaging; microstrip antennas; gain

1. INTRODUCTION

Recently, microwave imaging has received a considerable amount of interest with respect to the task of detection and location of malignant tissue in the woman’s breast [1–6]. A microwave imaging system is considered as a viable alternative to X-ray mammography due to its several advantages such as cost and insignificant side-effects. Microwave imaging involves the propagation of very low levels (1000 times less than a mobile phone) of microwave energy through the breast tissue. The basis for tumor detection and location is the difference in the electrical properties of normal and malignant breast tissue. Normal breast tissue is largely transparent to microwave radiation, while the malignant lump contains more water and blood, resulting in microwave signal back scattering. This scattered signal can be picked up by a microwave antenna and can be analyzed using a computer [5].

In general, two approaches are used with respect to detecting cancerous tissue. In one approach, known as microwave tomography [1–3], a forward and reverse electromagnetic field problem is solved to detect and locate a tumor in women’s breast. Each of these problems is solved at a single frequency; however, it has been found that a multiple-frequency approach enhances the detection process [3].

An alternative approach is microwave imaging, which involves generating and receiving short pulses for the various locations of a probe antenna or alternatively by an array antenna [4–6]. Such short pulses can be generated in practice by applying a step-frequency pulse synthesis technique [7]. The space or time-domain representation is then achieved using an inverse fast Fourier transform. The processed signals for the various locations of a probe antenna or from array elements are combined to form a two- or three-dimensional (3D) image showing the location of highly reflecting objects representing a cancerous tissue [7].

The multiple-frequency tomography approach and the radar technique require the use of ultra-wideband (UWB) antenna elements preferably in the planar format. Many designs of UWB antennas have recently been reported with respect to UWB communications. Most of these antennas represent some form of planar monopole, and they include rectangular and circular shapes [5–10]. An alternative UWB planar antenna element, which finds use in radar applications, is a unipolar or antipodal Vivaldi antenna [11].

The present designs of these antennas do not make them straight-forwardly applicable to microwave imaging applications. The reasons are as follows: UWB planar monopoles for use in UWB communication applications feature an omnidirectional radiation pattern and low gain. If one attempts to use them in a microwave imaging system, the dynamic range is sacrificed. In turn, Vivaldi antennas designed for radar applications usually exhibit high directivity and this is accomplished by using large physical size. Such large antennas are difficult to accommodate in a microwave imaging system.

In the recently reported microwave imaging systems for breast cancer detection, resistively loaded monopoles [3, 5] and miniature pyramidal horn antennas [5, 6] have been used. The shortcoming of the resistively loaded monopoles is that in addition to having inherently low directivity (they do not focus their beam), they incur power losses, which adversely affect the dynamic range of the microwave instrumentation. In turn, the miniature pyramidal horns, which are capable of focusing power on the imaged object, are elaborate to manufacture.

This article addresses the shortcomings of the present generation of antennas used in UWB imaging systems by designing a compact planar UWB antenna element of moderate gain. The proposed antenna element is in the form of a planar tapered slot made of a high dielectric constant substrate material to achieve its compact size. It features a very low loss across the desired band, and its radiation efficiency exceeds 90% with a relatively high gain. The design of UWB antenna as proposed in this article is accomplished using simple design formulas. This is an advantage with the previously reported UWB antenna designs [8–11] relying on a trial and error method and simulation tools. The presented design offers a compact antenna size, the topic not addressed in [11] while designing Vivaldi type antennas.

The paper is organized as follows. Section 2 describes the configuration of the circular cylindrical microwave imaging system. Section 3 describes the design procedure of an UWB antenna for inclusion in this array. Section 4 shows the performance data of the designed antenna obtained by both simulation using Ansoft HFSS and measurements. Finally, Section 5 concludes the article.

2. PROPOSED CIRCULAR CYLINDRICAL ANTENNA ARRAY

Figure 1 shows the configuration of a microwave imaging system including a circular array of UWB planar antenna elements. In this system, one of the antennas is used to transmit a microwave signal, while the rest of the antennas in the array receive the scattered signal. The measured data are collected and then the measurement procedure is repeated with the second antenna transmitting the signal, while the remaining ones are used for receiving the scattered signal. This process is continued until all antennas in the array perform the transmitting role. Note that the antenna array can be moved up and down automatically via a computer-controlled high-precision linear actuator to facilitate the collection of data for creating a 3D object image.
3. ANTENNA DESIGN METHODOLOGY

The configuration of a planar tapered slot antenna which is aimed for inclusion in an UWB microwave imaging system is shown in Figure 2. It is in the form of antipodal Vivaldi antenna [11]. The design objective is to obtain a compact size while preserving operation over the bandwidth of 3.1–10.6 GHz. This is the allowable frequency band for UWB applications [12]. To reduce the time and effort needed in the trial-and-error strategy adopted by other papers, we propose a simple design procedure of this antenna, whose validity is confirmed by EM simulations and measurements. The design steps are as follows.

Step 1: Given the lowest frequency of operation ($f_1$), thickness of the substrate ($h$) and its dielectric constant ($\varepsilon_r$), the width ($w$) and length ($l$) of the antenna structure excluding the feeder can be calculated using Eqs. (1) and (2).

$$w = l = \frac{c}{f_1} \sqrt{\frac{2}{\varepsilon_r + 1}}, \quad (1)$$

where $c$ is the speed of light in free space.

Step 2: The first radiating structure of the antenna is formed from the intersection of quarters of two ellipses. The major radii ($r_1$ and $r_2$) and the secondary radii ($r_{s1}$ and $r_{s2}$) of the two ellipses are chosen according to the following equations

$$r_1 = \frac{w}{2}, \quad (2)$$
$$r_2 = \frac{w}{2} - w_m, \quad (3)$$
$$r_{s1} = l - a, \quad (4)$$

$$r_{s2} = 0.38 r_2. \quad (5)$$

The parameter $a$ is used to control the lowest frequency of operation.

Step 3: The width of the microstrip transmission feeder $w_m$ to give the characteristic impedance, $Z_0 = 50 \, \Omega$, can be calculated using the following equations [13]:

For $w_m/h \leq 1$

$$Z_0 = \frac{60}{\sqrt{\varepsilon_{me}} \ln \left( \frac{8h}{w_m} + \frac{w_m}{4h} \right)}, \quad (6a)$$

and for $w_m/h \geq 1$

$$Z_0 = \frac{120\pi}{\sqrt{\varepsilon_{me}}} \frac{1}{w_m/h + 1.39 + 0.67 \ln(w_m/h + 1.44)}, \quad (6b)$$

where the effective dielectric constant for the transmission line, $\varepsilon_{me}$ is given by

$$\varepsilon_{me} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2 \sqrt{1 + 12h/w_m}}. \quad (7)$$

Step 4: The part including a ground plane of the antenna is similar to that of the first radiating element of the antenna. This part is a tapered-structure formed from the intersection of a rectangular conductor with two antifaced quarter ellipses with dimensions $r_2$ and $r_{s2}$. To improve the impedance matching of the antenna, the ground plane is extended by $y_g$.

Figure 2  Configuration of the proposed planar tapered slot antenna. (Parameters: $w = 59.6 \, \text{mm}$, $l_d = 59.9 \, \text{mm}$, $w_t = 0.5 \, \text{mm}$, $y_t = 12.9 \, \text{mm}$, $r_1 = 29.8 \, \text{mm}$, $r_2 = 29.3 \, \text{mm}$, $r_{s1} = 1.67$, $r_{s2} = 0.38$, $y_g = 1.8 \, \text{mm}$, $h = 0.635$, and $a = 0.28$). [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
4. ANTENNA PERFORMANCE

The validity of the presented design is tested by designing an antenna covering the UWB frequency band from 3.1 to 10.6 GHz. The design assumes Rogers RT6010LM substrate, featuring a dielectric constant of 10.2 and a loss tangent of 0.0023, 0.64-mm thickness plus 17-μm-thick conductive coating. The photograph of the manufactured UWB antenna, which was designed using the above outlined design procedure, is shown in Figure 3. As can be seen in Figure 3, the antenna is of a compact size of 50 × 50 mm².

The return loss and radiation pattern of the designed antenna is first verified using a finite element method design and analysis package Ansoft HFSS v 9.2. A personal computer with dual Xeon 2.8-GHz processors and 3.5 GB of RAM is used as a simulation platform. The developed antenna return losses and radiation pattern are tested in an anechoic chamber using an HP8530/HP8510 receiver/network analyzer.

Figure 4 shows the simulated and measured return loss of the proposed planar tapered slot antenna. As can be seen from the figure, the antenna operates from 2.75 GHz to over 11 GHz. The measured return loss graph closely resembles the simulated result, which confirms the validity of the design. The far field radiation patterns of the antenna in the two principle planes, xz plane (φ = 0) and the yz plane (φ = 90) are shown in Figure 5 at three different frequencies. The antenna patterns were measured at 3, 6, and 9 GHz and the results are shown in Figures 5(a)–5(c), respectively. From the figure, it can be seen that at the yz plane, directivity can be observed for the three measured frequencies. The front-to-back ratio is at least greater than 11 dB for the three different measured frequencies, hence, demonstrating the directive properties of the antenna needed for the proposed imaging system. The variation of gain with frequency of the antenna is shown in Figure 6. The simulated and measured gain for the antenna for...
frequencies between 3 and 10 GHz shows that the minimum gain is at 3 GHz and its maximum at 7 GHz with a gain of 9.4 dBi.

The last test concerns the ability of the manufactured antenna to transmit and receive pulses without distortions. In this case, two identical antennas are used to measure the transmission coefficient between the two antenna ports in the frequency domain and these results are transformed (via an inverse Fourier transform) to the time domain using time-domain capability of HP8510C/HP8530 VNA/receiver. The results for the transmitted and received pulse are shown in Figure 7, when the two copolarized antennas are separated by the distance of 45 cm. In this figure, the received pulse is scaled, so that its peak value is the same as that of the transmitted pulse. It can be seen in Figure 7 that the 3-dB width of the received pulse is equal to that of the original pulse. The pulse distortion is observed at the magnitude less than 0.1 with respect to the peak value. This result indicates that the developed antenna supports narrow pulses almost without distortions, as required in an UWB imaging system.

5. CONCLUSION

In this article, a method for designing an UWB planar tapered slot antenna for use in a circular cylindrical microwave imaging system has been presented. Based on the presented method, a compact antenna with an extremely wide 10 dB return loss bandwidth from 2.75 GHz to more than 11 GHz has been designed and developed. This antenna element features directive properties with gain ranging from 3.5 to 9.4 dBi over the required ultra-wideband. The antenna with an extremely wide 10 dB return loss bandwidth from 3.5 to 9.4 dBi over the required ultra-wideband. The last test concerns the ability of the manufactured antenna to transmit and receive pulses without distortions. In this case, two identical antennas are used to measure the transmission coefficient between the two antenna ports in the frequency domain and these results are transformed (via an inverse Fourier transform) to the time domain using time-domain capability of HP8510C/HP8530 VNA/receiver. The results for the transmitted and received pulse are shown in Figure 7, when the two copolarized antennas are separated by the distance of 45 cm. In this figure, the received pulse is scaled, so that its peak value is the same as that of the transmitted pulse. It can be seen in Figure 7 that the 3-dB width of the received pulse is equal to that of the original pulse. The pulse distortion is observed at the magnitude less than 0.1 with respect to the peak value. This result indicates that the developed antenna supports narrow pulses almost without distortions, as required in an UWB imaging system.

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Design of Compact Directive Ultra Wideband Antipodal Antenna

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Abstract: In this paper, a novel design procedure for designing a compact UWB antipodal Vivaldi antenna is presented. The antenna operates over the UWB frequency band from 3.1 to more than 10.6 GHz. Its measured far-field radiation is directive and its peak gain is 10.2 dBi in the specified band. The antenna pulse response shows negligible distortion, indicating that it can be useful in a precision ranging and imaging instrumentation. © 2006 Wiley Periodicals, Inc. Microwave Opt Technol Lett 48: 2448–2450, 2006; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21955

Key words: microstrip antenna; UWB; broadband

1. Introduction

Ultra-wide band (UWB) antennas have received a significant amount of attention since 2002, due to the approval by the Federal Communication Commission (FCC) of a 3.1 to 10.6 GHz frequency band for use in UWB communication, measurement, and radar systems [1]. The antenna is one of the essential components of these systems. Because of the requirement of small size transceivers, it is often required to have a compact size. The majority of the compact UWB antennas presented in the literature exhibit omnidirectional radiation patterns with relatively low gain and an impulse response with observable distortion [2]. These types of UWB antennas are suitable for short range indoor and outdoor communication; however, for radar systems such as an UWB microwave imaging system for detection of tumor in woman’s breast [3], moderate gain directive antenna is advantageous. In addition to an UWB impedance bandwidth, as defined by the minimum return loss of 10 dB, this antenna is required to support the subnanosecond pulse transmission with negligible distortion. This is necessary to achieve precision imaging without ghost targets. The unipolar and antipodal Vivaldi antennas presented in the literature [4, 5] satisfies the requirements for imaging systems in terms of bandwidth, gain, and impulse response albeit at the expense of significant volumetric size. Therefore, the challenge is to reduce their physical dimensions such that it can be incorporated in a compact microwave imaging detection system whilst maintaining its distortionless performance.

In this paper, the design of a compact antipodal Vivaldi antenna, which meets the above-mentioned requirements, is presented (Fig. 1). The antenna is fabricated on a high dielectric constant material, Rogers RT6010LM (εr = 10.2, h = 0.64 mm). The antenna is designed using simple design procedures resulting into compact antenna, which are verified using commercial software package and experimental tests. The simulation and experimental results show that the antenna covers the required UWB band of 3.1 to 10.6 GHz and features directive radiation properties. The duration of the pulse response of the antenna is <1 ns with negligible distortion making it an excellent candidate for a compact microwave imaging system.

2. Configuration and Design Strategy

The proposed antipodal Vivaldi antenna for inclusions in an UWB microwave imaging system [3] is shown in Figure 2. The design objective is to obtain its compact size whilst maintaining the bandwidth requirement of 3.1–10.6 GHz. The following design procedure is proposed and utilized in developing the proposed antipodal Vivaldi antenna:

2.1. Step 1

Given the lowest frequency of operation (f1), thickness of the substrate (h), and its dielectric constant (εr), the width (w) and length (l) of the antenna structure, excluding the feeder can be calculated using Eqs. (1) and (2).

\[ w = l = \sqrt{\frac{2}{f_1 \sqrt{\varepsilon_r + 1}}} \]  

where \( c \) is the speed of light in free space.

2.2. Step 2

The radiating structure of the antenna is formed from the intersection of quarters of two ellipses. The major radii (r1 and r2) and the secondary radii (r1s and r2s) of the two ellipses are chosen according to the following equations:

\[ r_1 = \frac{w}{2} + \frac{w_d}{2}, \]  

\[ r_2 = \frac{w}{2} - \frac{w_d}{2}, \]  

\[ r_{1s} = l + a, \]  

\[ r_{2s} = 0.642, \]

The parameter a is used to control the lowest frequency of operation.
2.3. Step 3
The width of the microstrip transmission feeder $W_m$ to give the characteristic impedance, $Z_0 = 50 \Omega$ can be calculated using the following equations [6]:

$$ w_m = \frac{120\pi h}{\sqrt{\varepsilon_f}} Z_0 $$  \hspace{1cm} (6)

The validity of the proposed design methodology is verified using the commercial software package, Ansoft HFSS, and experimental tests.

3. RESULTS
Figure 2 shows the simulated and measured return loss of the compact ($5.2 \times 5.2$ cm$^2$) antipodal Vivaldi antenna developed in Rogers RT6010LM ($\varepsilon_r = 10.2$, $h = 0.64$ mm) material. As shown in Figure 2, the 10 dB return loss bandwidth extends from 3 to more than 11 GHz covering the required UWB band of 3.1–10.6 GHz. The simulated result closely resembles the measured result validating the design procedure of the antenna. The far-field radiation patterns were measured at the two principal planes, $x$–$y$ and $y$–$z$ at 3, 6, and 9 GHz in an anechoic chamber. For brevity, only the 3 and 6 GHz patterns are presented and are shown in Figures 3(a) and 3(b), respectively. The antenna shows directive properties in the $y$–$z$ plane (note that broadside is $90^\circ$ on the plots) for both the measured frequency, with front-to-back ratio $>16$ dB, making it an ideal candidate for microwave imaging applications. The gain of the antenna is shown in Figure 4. The figure reveals a similar trend between the simulated and measured gain of the antenna. The measured peak gain is 10.2 dBi and that occurs at 8 GHz. Finally, the pulse response of the proposed antenna is tested. Here, two copolarized antennas are separated by the distance of...
45 cm and the results of the transmitted and received pulse are shown in Figure 5. Note that the two pulses are normalized with respect to their peak values. The figure reveals that the pulse duration of the antenna is 0.8 ns. The pulse distortion occurs at the 0.12 level with respect to the peak level of 1, and thus is almost negligible. The observed results indicate that the developed antenna supports narrow pulse almost without distortion making it an excellent radiator for the purpose of microwave imaging with high resolution.

4. CONCLUSIONS

A novel design method has been proposed to achieve a compact antipodal Vivaldi antenna. The designed and developed antenna operates from 3.1 GHz to more than 10.6 GHz and has directive radiation properties. The front-to-back ratio of the antenna is >16 dB and its peak gain is 10.2 dBi. The duration of the impulse response of the antenna is <1 ns with negligible distortion as required in precision ranging and imaging applications.

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DESIGN OF A COMPACT ULTRA-WIDEBAND ANTENNA
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ABSTRACT: A simple method is presented for the design of ultra-wideband antennas in planar format. This method is demonstrated for a high-dielectric-constant substrate material, which allows for a considerable antenna size reduction. Simulations are performed using Ansoft’s High-Frequency Structure Simulator (HFSS) for antennas assuming DuPont951 (εr = 7.8) and RT6010LM (εr = 10.2) substrates. For the 1-mm-thick DuPont951, the designed antenna with 22 × 28 mm dimensions features a 10-dB return loss bandwidth from 2.7 GHz to more than 15 GHz; for the 0.64-mm-thick RT6010LM a 20 × 26 mm antenna exhibits a 10-dB return loss bandwidth from 3.1 to 15 GHz. Both antennas feature nearly omnidirectional properties across the whole 10-dB return loss bandwidth. The validity of the presented UWB antenna design strategy is confirmed by measurements performed on a prototype developed on RT6010LM substrate. © 2006 Wiley Periodicals, Inc. Microwave Opt Technol Lett 48: 1515–1518, 2006; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21705

Key words: planar antenna; RF front end; ultra-wideband antenna

1. INTRODUCTION
Recent years have witnessed an increased interest in ultra-wideband (UWB) antennas since the adoption of UWB technology by US-FCC in 2002 [1]. In parallel to interest in UWB antennas, recent research has also focused on high dielectric multilayer ceramics, such as low temperature co-fired ceramics (LTCC), to reduce the size of front end modules in wireless transceivers [2]. Merging the two technologies is a challenging task because integration requires the development of planar UWB antennas on a high-dielectric-constant material.

Several methods have been proposed recently in order to realize a planar UWB antenna with suitable radiation characteristics. Examples include a planar volcano-smoke slot antenna [3, 4], a coplanar waveguide fed bowtie/triangular patch antenna [5], a multistructure coplanar waveguide (CPW)-fed [6], and planar monopole antennas [7]. The main drawback for the abovementioned designs is that they use a trial-and-error method with the help of a simulation tool to get the desired response.

In this paper, a simple method is presented for the design of UWB antenna. The proposed method gives a compact design with UWB characteristics. The technique introduced in this paper uses the intersection of elliptical structures to form the radiating element and the coplanar waveguide feeder. Steps for the design are presented in section 2. Section 3 presents results of the simulations obtained with Ansoft HFSS while section 4 presents results of measurements on an UWB antenna designed using the described method. Section 5 concludes the paper.

2. DESIGN
The configuration of the UWB antenna, which is the subject of investigations of this paper, is illustrated in Figure 1. The radiating crusade shaped slot is the result of intersection of two ellipses. The antenna is assumed to be fed using CPW to enhance its broadband characteristics. The steps used to design this antenna are summarized as follows:

In step 1, depending on the lowest frequency f1 of operation, thickness h of the substrate, and dielectric constant εr, the width w and length l of the antenna structure are calculated from [8] as

\[ \begin{align*}
w &= \frac{c}{2f_1 \sqrt{\varepsilon_r + 1}}, \\
l &= \frac{c}{2f_1 \sqrt{\varepsilon_r}} - 2\Delta l,
\end{align*} \]

Figure 1 Configuration of the ultra-wideband antenna. Different design parameters are shown (left) and the feeding structure (right). [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
In step 2, the radiating slot is formed by cutting an ellipse (ellipse 1) from the conductive layer covering the substrate designed above and adding a second ellipse (ellipse 2) to the ground plane in the manner shown in Figure 1. The major diameters $D_1$ and $D_2$ of ellipses 1 and 2, respectively, and ratios $R_1$ and $R_2$ of the secondary diameter to the major diameter in the two ellipses, are equal to:

\[ D_1 = \frac{C}{2\sqrt{\varepsilon_a}} \]  
\[ D_2 = \frac{C}{4\sqrt{\varepsilon_a}} \]  
\[ R_1 = R_2 = \frac{w}{T} = \frac{2\varepsilon_a}{\varepsilon_a + 1} \]  
\[ e_{\infty} = \frac{e_r + 1}{2}, \quad e_{\infty} = \frac{1 - e_r}{2} \left[ 1 + 12 \frac{h}{w} \right]^{-0.5} \]  
\[ \Delta l = 0.412h \left( \frac{e_{\infty} + 0.3)(0.264 + w/h)}{e_{\infty} - 0.258)(0.8 + w/h)} \right) \]  

where $c$ is speed of light in free space and $e_{\infty}$ is the effective dielectric constant. The antenna structure is assumed to be in the $x$-$y$ plane with its higher dimension extending along the $y$-axis. On top of the substrate, a conductive layer is assumed. This layer is used to form the radiating slot and the CPW as explained in the following steps.

In step 3, the centers of the two ellipses are shifted by $y_1$ and $y_2$ from center of the ground plane, as in Figure 1. Parts of the two ellipses that extend outside the ground plane from one direction are cut to form the CPW needed to feed the antenna. The values of $y_1$ and $y_2$ are chosen in order to maintain the required impedance bandwidth.

### 3. SIMULATION RESULTS

The UWB antenna was designed assuming two high-dielectric-constant substrates: one is an LTCC type, DuPont951, with a dielectric constant equal to 7.8, tangent loss (tan $\delta = 0.0015$) and thickness 1 mm, and the other is Rogers RT6010LM with a dielectric constant equal to 10.2, a tangent loss of 0.0023, and thickness of 0.64 mm. The design was tested using Ansoft HFSSv9.2. Concerning the conductive layers, a 17-$\mu$m thickness of copper metallization was assumed. The lowest frequency of operation was set to 2.5 GHz.

Table 1 shows dimensions of the designed antennas generated using Eqs. (1)–(7). Figure 2 shows variations of the return loss (RL) versus frequency for the designed antennas using DuPont951 and RT6010LM substrates. Dimensions of the antenna employing DuPont951 are equal to 22 $\times$ 28 mm while the dimensions are 20 $\times$ 26 mm for the antenna employing RT6010LM substrate. The computed characteristics of the designed antennas reveal UWB behavior with bandwidth from 2.7 GHz to more than 15 GHz for the DuPont951 and from 3.1 to 14.4 GHz for the RT6010LM assuming a 10-dB return-loss reference.

From the UWB applications point of view, the antenna is usually required to have an omnidirectional radiation. The designed antenna fulfills this requirement, as shown in Figure 3, for the antenna using DuPont951 substrate. Similar results were obtained for the antenna designed using RT6010LM and therefore these results are not shown here. Concerning gain, our calculations have shown that the gain (in dB) increases approximately linearly with frequency in the frequency band from 3 to 12 GHz and is between 1 to 6 dB for the two antennas. Then it reaches a plateau and slightly drops in value for the remaining frequencies in the 12–15-GHz band.

The final verification concerns variations of the radiation efficiency for the designed antenna. The results of our simulations have revealed that the investigated antennas feature high efficiency, being greater than 90% (and improving with frequency) in the 3–15-GHz band.

### 4. MEASUREMENTS

In order to verify the validity of the proposed design method, as well as to check its HFSS simulated performance, we manufactured and experimentally tested the second UWB antenna (the one employing RT6010LM).

The return loss and the radiation pattern of the antenna were measured using an HP6510/HP8530 network analyzer/receiver in an anechoic chamber. The return loss for the antenna is shown in Figure 4. It is clear that the antenna has UWB characteristics with a bandwidth that extends from 3.1 to 14.8 GHz assuming a 10-dB return-loss reference. The result is in a relatively good agreement with the simulated RL plot of Figure 2. The far field radiation pattern of the antenna in the two principle planes, $x$–$z$ plane ($\phi = 0$) and the $y$–$z$ plane ($\phi = 90$) are shown in Figure 5 at different frequencies. Nearly omnidirectional behavior, especially in the

![Figure 2](http://example.com/figure2.png)

**Figure 2** Return loss vs. frequency for the designed antennas obtained from Ansoft HFSS simulations. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
Variation of gain of the antenna with frequency was also measured. The result is shown in Figure 6. The measured gain is between 0.8 and 4 dB for the frequency range 3.1 to 15 GHz, and is slightly lower than simulated one. The measured gain curve shows similarities with the simulated one.

5. CONCLUSION

In this paper, a simple method to design a compact planar UWB antenna has been presented. The proposed radiating element is formed by the intersection of two ellipses, while the feeding structure is a coplanar waveguide. The designed antennas using this method have an extremely wide 10-dB return-loss bandwidth from around 3 to 15 GHz. They feature almost omnidirectional radiation patterns and exhibit more than 90% radiation efficiency. The presented simulated and measured results provide high confidence in the validity of the proposed design method of an UWB antenna.

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Gain and Bandwidth Optimization of Compact UWB Tapered Slot Antennas

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Abstract – A method is presented which is aimed at increasing the gain and bandwidth of the tapered slot antenna by just finding the optimum tapering profile of the antenna without the need to increase its size. The result of the analysis indicates that a slightly negative tapering gives the optimum gain and bandwidth performance. A positive tapering causes a sharp reduction in the gain and bandwidth, assuming a constant width and length for the antenna. The result of analysis is utilized by designing and manufacturing a tapered slot antenna which has an ultra wideband performance (bandwidth from 3.1 GHz to 10.6 GHz) and a maximum possible gain. The dimension of the antenna is 65 mm × 65 mm. The simulated and measured results show that the antenna covers the 3.1 GHz to more than 11 GHz band with a gain which varies between 4.5 dBi at 3 GHz and 12 dBi at the high frequency band (9-11 GHz).

Index Terms-Tapered slot antenna, ultra wideband.

I. INTRODUCTION

Recently, ultra wideband (UWB) systems have received a considerable amount of interest with respect to communication and medical applications. An antenna, which can efficiently radiate and receive UWB signals, is essential for successful operation of these systems. In some applications such as the biomedical imaging, high efficiency, light weight, compact size, and end-fire radiation characteristics are the important requirements that are required for these antennas [1]. The tapered slot antenna (TSA) is one possible candidate to meet such requirements. TSA has already been utilized in UWB radar. In this case, the TSA featuring a relatively high gain of 10-15dB is used. Such requirement translates into the large length of the antenna being several wavelengths at the centre frequency of a given band [2]. This design is unsuitable for biomedical applications, as they require the antenna to be of compact size.

In the present work, the possibility of increasing the gain of a compact size TSA antenna by changing the tapering profile is investigated. The effects of the tapering profile on the beamwidth, sidelobe level and radiation pattern have already been extensively studied but no specific recommendations with respect to the gain are given [2-6]. The following investigations are concentrated on the TSA covering the UWB, i.e. 3.1-10.6 GHz.

II. ANALYSIS

In order to study effect of the tapering profile on the gain and bandwidth of the tapered slot antenna an exponential tapering shape was assumed. The value of the exponent is to be changed so that a wide range of tapering profile can be covered. To this purpose, a tapering parameter $B$ is introduced. The analysis includes the following steps:

Step 1: The length and width of the radiating element, which is the tapered slot, are chosen to be equal to the effective wavelength calculated at the lowest frequency of operation. Given the lowest frequency of operation ($f_l$), thickness of the substrate ($h$) and its dielectric constant ($\varepsilon_r$), the width ($w$) and length ($l$) of the antenna...
The TSA design was assessed for different values of $B$ assuming Rogers RO4003 ($\varepsilon_r = 3.38$, thickness=0.508mm) as a substrate. This task is performed using the frequency domain finite element method. The calculated values of the average gain and bandwidth is shown in Fig.2. In this figure the required 7.5GHz bandwidth for UWB applications is shown. The result indicates that the antenna has a maximum gain (11.7 dBi) when $B=0$. In turn, a maximum bandwidth is obtained when $B=-0.25$. For a positive tapering ($B$ is positive), the gain and the bandwidth decrease rapidly. To compromise between the highest possible gain and the required BW (7.5 GHz in the present case), the optimum value for the parameter $B$ is chosen as -0.12 (this is the intersection of the two graphs). 

III. RESULTS

In order to develop and test a tapered slot antenna using the optimum value of $B$, it is preferred to alter the feeding structure of the antenna so that a microstrip feeder can be used instead of the slot line. It is well known that the microstrip line is the most common form of printed transmission line used for feeding a tapered slot antenna element. Therefore, the developed antenna was modified in order to use a microstrip line as a feeder. A microstrip line, being formed by a conductive metal strip on one side of a dielectric substrate and a conductive ground plate on other side of the substrate, is an unbalanced line [7]. This is opposite to the slot line, which is a balanced transmission line. Because of this situation, feeding a TSA with a microstrip line
requires a wideband balanced-to-unbalanced transition (balun) to avoid compromising the broadband performance.

Different methods for the microstrip line feeding arrangement for the TSA were presented in [7]-[9]. In this paper, the following arrangement was used. The slot line was terminated in an open circuit by adding a relatively large circular patch at the end of the slot line, and the microstrip line was shorted by using a circular patch, which effectively acts as a shorting via. Due to the above mentioned arrangement, a balun is created at the crossover which matches the unbalanced microstrip line to the balanced slot line of the antenna element. This electromagnetic coupling arrangement permits signal transmission from the microstrip transmission line to the slot line (for feeding the antenna). In general, the stronger the electromagnetic coupling, the better is the transition. Fig. 3 shows the type of balun which was used for the developed antenna. This configuration has no inherent bandwidth limitation other than parasitic inductances and capacitances.

The validity of the presented analysis was tested by designing an antenna with $B=-0.12$ in order to cover the UWB frequency band from 3.1GHz to 10.6GHz using Rogers RO4003C as a substrate. Dimension of the developed antenna is 65 mm $\times$ 65 mm. A photo for the manufactured antenna is shown in Fig.4.

![Fig.3 Configuration of the microstrip/slot line transition as a method to feed the antenna using a microstrip line.](image)

![Fig.4 The manufactured antenna. (a) Top layer revealing the radiator and the transition, and (b) bottom layer showing the microstrip feeder.](image)

Fig. 5 shows the simulated and measured return loss of the manufactured planar tapered slot antenna. As can be seen from Fig.5, the 10dB return loss of this antenna extends from 3.1GHz to over 11GHz. This result agrees well with the result of calculation shown in Fig.2, where the 10 dB bandwidth was expected to be more than 7.5 GHz when $B=-0.12$.

![Fig.5 Variation of the measured and simulated return loss with frequency.](image)

The measured far-field radiation patterns of the antenna in the two principle planes are shown in Fig.6 at three frequencies (4, 7 and 10 GHz). The measured patterns reveal that the front-to-back ratio of the antenna is greater than 12 dB indicating directive properties.

![Fig.6 The measured far-field radiation patterns.](image)

The variation of the measured and simulated gain with frequency is shown in Fig.7. The simulated gain for the antenna for the frequencies between 3 and 11GHz shows that the gain increases with frequency and is around 13 dBi at 11GHz. The measured gain of the antenna also increases with
frequency. It is between 4.5 dBi at 3 GHz and 12 dBi at the frequency band 9 GHz to 11 GHz. The simulated and measured results agree with each other, and both of them agree with the expected value shown in Fig. 2 for $B = 0.12$.

Fig. 6 The measured radiation pattern three frequencies.

Fig. 7 Variation of the measured and simulated gain with frequency.

IV. CONCLUSION

A method has been presented which aimed at increasing the gain and bandwidth of the tapered slot antenna by just changing the tapering profile of the antenna without the need to increase its size. The result of analysis has indicated that a slightly negative tapering is the optimum choice for the tapered slot antenna with ultra wideband performance and compact size. The method has been utilized by designing and manufacturing a tapered slot antenna which has an ultra wideband performance (bandwidth from 3.1 GHz to 10.6 GHz) and a maximum possible gain. The simulated and measured results show that the antenna covers the 3.1 GHz to more than 11 GHz band with a gain which varies between 4.5 dBi at 3 GHz and 12 dBi at the high frequency band (9-11 GHz).

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An Ultra Wideband Microwave Imaging System for Breast Cancer Detection

Wee Chang Khor†, Marek E. Bialkowski†, Amin Abbosh†, Norhudah Seman† and Stuart Crozier†

Summary
An experimental study concerning Ultra Wideband (UWB) Microwave Radar for breast cancer detection is described. A simple phantom, consisting of a cylindrical plastic container with a low dielectric constant material imitating fatty tissues and a high dielectric constant object emulating tumour, is scanned with a tapered slot probe antenna operating between 3.1 to 10.6 GHz. A successful detection of a target is accomplished by a visual inspection of a two-dimensional image of the scanned phantom.

Key words:
Cancer, Microwave Imaging, Antennas, Biomedical Imaging.

1. Introduction
For years, the detection of breast cancer has relied on film (X-ray) mammography, and as a result it is considered as the “gold standard” for breast cancer diagnosis. Despite this fact, it still features a number of shortcomings. For example, it is ineffective for women with dense breasts. Moreover, as it involves radiation it introduces health risks in the case of frequently made tests [1].

An ultimate diagnosis of all types of breast disease depends on a biopsy. A biopsy is an invasive procedure to remove and examine tissue or cells for the presence of cancer. In most cases the decision for a biopsy is based on mammography findings. Unfortunately, biopsy results indicate that 80% of breast lesions detected by mammography are benign [2]. This situation calls for alternative diagnostic tools to reduce physical and mental suffering of patients caused by this false positive diagnosis. Of particular interest is the development of low-cost diagnosis methods, which could be easily accessed by the masses.

Recent research in breast cancer detection concentrates on such alternatives as magnetic resonance imaging, ultrasound tomography, microwave tomography and microwave radar [3].

Magnetic Resonance Imaging has recently been shown to be a very useful screening tool [4], but is expensive and time consuming [2]. MRI also lacks the ability to image calcifications which are tiny calcium deposits that can indicate early breast cancers [5].

Ultrasound imaging of the breast is capable of distinguishing between solid tumours and fluid-filled cysts. Also, it can be used to evaluate lumps that are hard to see on a mammogram. As ultrasound does not harm biological tissue, thus it can be applied frequently. This is of importance, especially with respect to younger women for whom the risks from X-ray radiation are most significant. However, ultrasound lacks spatial resolution, cannot image calcifications and is very operator dependent [6].

Microwave techniques involve the propagation of very low levels (1000 times less than a mobile phone) of microwave energy through the breast tissue. The basis for tumour detection and location is the difference in the electrical properties of normal and malignant breast tissue. Normal breast tissue is largely transparent to microwave radiation while the malignant one containing more water and blood, causes microwave signal back scattering. This scattered signal can be picked by a microwave antenna and analysed using a computer [1], [3], [7-8].

In Microwave Tomography, a forward and reverse electromagnetic field problem is solved to detect and locate cancerous tissues in woman’s breast. Each of the inverse problems is solved at a single frequency [7] [8].

The radar approach to microwave imaging employs generating and receiving short pulses for various locations of probe antenna or alternatively by an array antenna [3] [9-10]. Such short pulses can be generated in practice by applying a step-frequency pulse synthesis technique [11]. The space or time-domain representation is then achieved using an Inverse Fast Fourier Transform (IFFT). The processed signals for various locations of a probe antenna or from array elements are combined to form a two or three-dimensional image showing the location of a highly reflecting object representing a cancerous tissue [12].

The first configuration, shown in Fig. 1(a), is based on the principle of monostatic radar [13], [14]. In this configuration, the same antenna is used for both transmitting and receiving of a microwave signal. As a result, the transceiver performs the function of a reflectometer [14], [15]. The configuration shown in Fig. 1(b) uses two antennas, which are displaced by some distance. In this case, the microwave imaging system is based on the principle of bistatic radar [13].
Based on the monostatic radar, the first generation linear/circular scanning system prototype, which was built at the University of Queensland, was described in [12]. The prototype using an open ended circular waveguide, employed as a probe antenna, and operating over a limited frequency band of 8.2 to 12.4 GHz was shown to be successful in detecting small targets featuring high conductivity. This paper reports on a significant extension of this initial prototype. The system is modified to be an UWB system covering the frequency range of 3.1 to 10.6 GHz. This operation is possible with the use of an UWB tapered slot antenna.

The paper is organised as follows. Section 2 describes the configuration of the UWB radar system. Section 3 describes the design of the Tapered Slot UWB antenna. Section 4 presents the results of experimental imaging of a circular container filled with vegetable oil which emulates the skin and healthy breast tissue respectively. A high dielectric constant object (a small plastic container filled with distilled water) is used to emulate the target tumour. Finally, Section V concludes the paper.

2. Experimental Setup

The configuration of the prototype UWB radar system is shown in the Fig.2.

Prior to measurements, the system is calibrated over the frequency band from 3.1GHz to 10.6GHz using a modified one-port calibration procedure that involves three broadband standard loads. The first two standard loads are the coaxial short and shielded open circuit, the same as in the standard calibration procedure. However for the third standard, the coaxial match termination is replaced by a load realized by the probe antenna radiating a microwave signal in free space [12]. We call this modified calibration procedure as Method A. By using Method A, undesired signals including internal reflections inside the probe antenna and at the antenna-air interface are either reduced or removed completely [12].

The calibration is followed by the measurement procedure, which includes the following steps. First, the area to be scanned is specified. The required information includes the
step size and number of steps for the Φ-Y scanning platforms. At each Φ-Y probe location, the PC controller triggers the source in the Vector Network Analyser and 50 to 800 (depending on specifications) measurement points for reflection coefficient over the frequency band of interest are performed. After the frequency domain measurements of reflection coefficients are completed, the obtained data is converted to the time domain using IFFT. Having obtained the frequency and time domain results for a given location, the results are stored in the PC and the probe is moved to a new position. Then, the measurement procedure is repeated. The obtained data is combined and then processed by the PC to create an image. Using false colours, the location of a target is shown in a colour distinctive from the colours representing other parts of the breast phantom.

3. Antenna Design

The design of the Tapered slot UWB antenna is accomplished using design formulas described in [16]. This design procedure is very simple, which forms an advantage over previously reported UWB antenna designs which rely on the trial and error method and sophisticated full EM analysis and simulation tools. Following its design, the antenna is fabricated on a Rogers RT6010LM substrate featuring a dielectric constant of 10.2 and a loss tangent of 0.0023, 0.64mm thickness plus 17μm thick conductive coating. This radiating element exhibits a compact size of 50mm × 50mm, which is important for the present application. The configuration and photograph of the antenna is shown in Fig.4 and Fig.5 [16].

The return loss performance of the designed antenna is first verified using a Finite Element Method design and analysis package, Ansoft HFSSv9.2. The developed antenna’s return losses are tested in an anechoic chamber using an HP8530/HP8510 microwave receiver/network analyser. Fig.6 shows the simulated and measured return loss of the planar tapered slot antenna. As can be seen from Fig.6, the antenna operates from 2.75 GHz to over 11 GHz for the 10dB return loss reference. The measured return loss graph closely resembles the simulated one. This antenna is definitely capable of operating in the desired frequency range from 3.1GHz to 10.6GHz [16]. This radiating element features good Front-to-Back ratio of 11dB meaning that it is able to mainly concentrate the radiated power on the breast phantom. This is of considerable advantage in comparison with omni-directional elements reported in other microwave breast radar systems (for example, as described in [10]).

4. Results and Discussion

The imaging capabilities of the above-described UWB radar system (Fig.7) are carried out for a simple breast phantom. The phantom consists of a circular cylindrical plastic container with a diameter of 12.5cm with thickness
of 1mm filled with vegetable oil (characterized by a relative dielectric constant of 4). The container and oil represent the skin layer and the breast tissue, respectively. A second small plastic container filled with water (relative dielectric constant of 80) representing the tumour is located inside the first plastic container. This experimental model represents quite well actual healthy breast tissues and cancerous tumours, which have typical relative dielectric constants of 9 and 50, respectively [9]. The present experimental setup is a considerable extension of our 1st generation Microwave imaging system, which was reported in [12].

![Image of experimental setup](image)

**Fig. 7: Experimental Setup**

Fig. 8 shows the imaging result obtained with the new system for a 30mm cylindrical water target scanned in a horizontal plane at angular increments of 22.5°.

![Image of imaging result](image)

**Fig. 8: Imaging Result of UWB Radar System**

The figure clearly shows the boundaries of the plastic container and the location, size and shape of the water target. This is a vast improvement over the 1st generation system which operated only over the 8.5-12.4GHz band. The use of a narrower frequency band in the old system resulted in poorer image resolution, which caused smearing and blurring of the target. As observed in Fig. 8, the new UWB Radar system better handles the resolution problem because it employs a larger frequency bandwidth. The image of the target is about 3.3cm in diameter which is very close to the actual physical diameter of 3cm. The next undertaken tests concern the capability of the UWB Radar system to detect small targets. Figure 12 and 13 shows the imaging result of cylindrical water targets with a diameter of 11mm and 5mm, respectively.

![Image of imaging result with calibration method A](image)

**Fig. 12: Imaging Result of UWB Radar System of a water target with a diameter of 11mm.**

![Image of imaging result with calibration method A](image)

**Fig. 13: Imaging Result of UWB Radar System of a water target with a diameter of 5mm.**

The presence and location of the two water targets can be clearly identified in the two figures. This means that by using the presented radar technique, targets as small as 5mm in diameter can be detected just by a simple visual inspection of the produced images. At present, we are carrying out further investigations into the use of various signal processing techniques to enhance
the presence of a target in the obtained radar image. One simple enhancement technique is via compensation of the received signal strength drop due to an increased distance between a target and a transmitting/receiving antenna. This topic is outside the scope of the work presented in this paper and thus is not covered here.

5. Conclusion

A UWB microwave imaging system based on a step frequency synthesised pulse technique for possible use in breast cancer detection has been presented. The system has been tested when an imaged object is a cylindrical plastic container filled with a low dielectric constant liquid and small size high dielectric constant cylindrical target. It has been found that the use of UWB signal enables detection of small targets in the order of a few millimetres. This can be accomplished by a visual inspection of an imaged produced by the developed UWB radar system.

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Compact coplanar waveguide-fed ultra-wideband antenna

H.K. Kan, W.S.T. Rowe and A.M. Abbosh

A novel coplanar waveguide fed planar antenna with an extremely broad bandwidth in excess of 128% is presented. This comfortably covers the required bandwidth for ultra-wideband communication applications, and exhibits the required omnidirectional pattern characteristics. The antenna is also very small in size, only 0.3 λ square at 3 GHz.

Introduction: Ultra-wideband (UWB) technology has received a considerable amount of attention since 2002. This interest stemmed from the Federal Communication Commission (FCC) granting commercial access to the frequency spectrum 3.1–10.6 GHz for use in imaging systems, communication and measurement systems, and vehicular radar systems. UWB communication systems for short-range high-speed indoor data communication applications require an antenna that is small in size with a wide bandwidth and omnidirectional radiation patterns. Many planar broadband antennas have been studied and reported for UWB applications that use a variety of antenna configurations, for example [1–3]. Each has its own merits and drawbacks; however, in most instances it is difficult to obtain an antenna that is small in size, yet still satisfies the bandwidth requirements for UWB applications. In some configurations, the ground-plane of the antenna is on the opposite side of the microstrip feed line, which necessitates the antenna to be processed on both sides of the circuit board. Coplanar waveguide (CPW) fed antennas [4–8] are a promising candidate for UWB systems for several reasons, such as ability to offer wide bandwidth, and that the antenna can be printed on a single side of the printed circuit board, hence alleviating the problem of space restrictions in a device. The majority of CPW-fed UWB antennas reported in the literature utilises a quasi-monopole/dipole architecture. This is due to the configuration being able to provide the omnidirection pattern and broad bandwidth required for UWB communications.

In this Letter, we present a compact ultra-wideband printed antenna. The antenna utilises a simple CPW feed, therefore alleviating the need for two sides of the printed circuit board to be etched connected to a half annular ring radiator. The 10 dB return loss of the antenna covers the required UWB spectrum and demonstrates an omnidirectional pattern at the co-polarised plane, satisfying the requirement for short range UWB communication.

Configuration and design strategy: The schematic of the proposed UWB antenna is shown in Fig. 1. As can be seen from the Figure, the antenna consists of a radiator shaped as half an annular ring. The outer radius (R_{out}) and inner radius (R_{in}) of the annular ring are 12 and 5 mm, respectively, with an opening ratio in the y direction of 1.2 for the outer and inner radius. The antenna is fed by a CPW transmission line and, as is evident from Fig. 1, the CPW ground plane is tapered away from the antenna elements, similar to those in [2] and [7]. The tapered ground plane in close proximity to the radiator is critical to achieve a wide bandwidth. The tapering of the ground plane can also minimise its extent, which assists in reducing

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**Fig. 1** Configuration of proposed CPW-fed ultra-wideband antenna

**Fig. 2** Return loss performance of antenna

**Fig. 3** Radiation patterns of antenna  
- a Simulated at 3.5 GHz  
- b Measured at 3.5 GHz  
- c Measured at 7 GHz
the overall antenna footprint. The resulting antenna structure is more balanced in nature than a standard CPW-fed antenna element. Hence, the antenna exhibits dipole-like behaviour, particularly with regard to the omnidirectional radiation pattern. This makes the proposed antenna well suited to UWB applications. The antenna is small in size, with overall dimensions $x_{sub}$ and $y_{sub}$ equal to 30 mm. The antenna is developed on a high dielectric constant Rogers Duroid 6010 substrate, which has a relative permittivity of 10.2. Ansoft HFSS was utilised in assisting to design the proposed antenna.

Conclusions: A novel coplanar waveguide fed planar antenna has been presented, which addresses the requirements for ultra-wideband communication applications. A 10 dB return loss bandwidth in excess of 128% is produced and an approximate omnidirectional pattern is radiated across the majority of this bandwidth. The antenna is compact, have a widest dimension of 30 mm.

References

Simple Broadband Planar CPW-Fed Quasi-Yagi Antenna

H. K. Kan, Member, IEEE, R. B. Waterhouse, Senior Member, IEEE, A. M. Abbosh, and M. E. Bialkowski, Fellow, IEEE

Abstract—In this letter, we present a novel coplanar waveguide fed quasi-Yagi antenna with broad bandwidth. The uniqueness of this design is due to its simple feed selection and despite this, its achievable bandwidth. The 10 dB return loss bandwidth of the antenna is 44% covering X-band. The antenna is realized on a high dielectric constant substrate and is compatible with microstrip circuitry and active devices. The gain of the antenna is 7.4 dBi, the front-to-back ratio is 15 dB and the nominal efficiency of the radiator is 95%.

Index Terms—Broadband antennas, coplanar waveguide feed, microstrip antennas, planar antenna, quasi-Yagi antenna.

I. INTRODUCTION

Planar quasi-Yagi antenna have received renewed interest recently due to its suitability for a wide range of application such as wireless communication systems, power combining, phased arrays, active arrays as well as millimeter-wave imaging arrays. Various designs of the planar quasi-Yagi antenna have been reported in the literature covering the X-band of the frequency spectrum [1]–[3]. In each of the design, the driver and the director element are similar, however, the most obvious distinction is the feeding mechanism employed. Typical feeding methods utilized include a microstrip feed [1], [2] or a coplanar waveguide (CPW) feed [3] each requiring a balun to transform the transmission line mode at the input port of the antenna to the coplanar stripline. Each of these has its own merits; however, the common major drawback is the feed of each requires a relatively complicated balun that not only increases the size of the structure, but can also degrade the radiation performance of the antenna. Recently, an attempt to alleviate this complication with a simplified feed and modified printed Yagi antenna was presented [4] and resulted in similar performance in terms of its bandwidth, radiation properties and gain to the previously reported configurations. Here, the feeding structure consisted of a microstrip line that transitions to a transmission line formed by two parallel strips. One side of the driver element is connected to the feed while the other side of the driver element is connected to the ground connected via a coplanar stripline. Although, this feeding technique and driver configuration does not require a balun, it necessitates two side of the substrate to be etched.

In this letter, we present a coplanar waveguide-fed quasi-Yagi antenna. The antenna utilizes a simple CPW feed, therefore, alleviating the complicated feeding network commonly required for the design of quasi-Yagi antenna. In addition, the proposed antenna is on a single layer and is very compact. The 10 dB return loss bandwidth of the antenna is 44% operating from 7.7 to 12 GHz covering X-band.

II. CONFIGURATION AND DESIGN STRATEGY

A schematic of the proposed antenna is shown in Fig. 1. As can be seen from the figure, the antenna consists of two director elements, a driven element and a ground plane acting as a reflector. The antenna is fed by a CPW transmission line and as can be seen from Fig. 1 there is no complicated balun for matching the driven element to the antenna feedline. As a starting point to designing the antenna, the length of the driven element should be around \(1.4\lambda_{ofr}\) while the lengths of the directors should be in the order of \(1.4\lambda_{ofr}\) according to the Yagi design principles [5]. Here, \(\lambda_{ofr}\) refers to the effective wavelength at the lowest frequency of operation. It is calculated assuming the following value for the effective dielectric constant of the substrate: \(\varepsilon_e = (\varepsilon_r + 1)/2\), where \(\varepsilon_r\) is the dielectric constant of the substrate. The 10 dB return loss bandwidth of the antenna is 44% covering X-band.
In order to develop a compact antenna we chose the value $\lambda_{dir}$. The design frequency for the proposed antenna is 8 GHz which results in the following initial values of the antenna parameters assuming the substrate to be Rogers RT6010 (0.64 mm, $\varepsilon_r = 2.2$): $L_{dir} = \infty$ mm, $s_{dir1} = s_{dir2} = 7.2$ mm, $s_{ref1} = s_{ref2} = 1.5$ mm. The commercial package, Ansoft HFSSv9.2 based on a 3-D full-wave finite element method (FEM) was then utilized in assisting to optimize the proposed antenna for wide bandwidth covering the X-band. The antenna was optimized after simulation and the following dimensions were selected: $L_s = 19.2$ mm, $W = 2.5$ mm, $L_{dir} = 3.7$ mm, $s_{dir1} = s_{dir2} = 0.8$ mm, $w_{dir1} = w_{dir2} = 0.5$ mm, $L_{ref1} = 11.5$ mm, $w_1 = 1$ mm, $L_1 = 7.4$ mm, $L_2 = 8.6$ mm, $s_{ref} = 9.9$ mm, and $s_{ef} = 5.4$ mm. In this optimized quasi-Yagi design, the antenna operates from 8 GHz to 12 GHz. As can be seen from Fig. 1, the novelty of the proposed quasi-Yagi antenna is its simplicity in its feed structure while still achieving its broad bandwidth and directive radiation properties. To the author’s knowledge, this is the simplest planar quasi-Yagi ever reported. As the distance between the reflector and the driven element $s_{ref}$ is decreased, the resonant frequency of the antenna is lowered albeit at the cost of reduced bandwidth. It is interesting to note that the distance $s_{ef}$ performs the task of a balun, therefore alleviating the need for the complicated balun commonly required. The total area of the substrate is approximately $|L_1 \times L_2|$ (where $\lambda_0$ is the free-space wavelength) at the central frequency. The antenna was manufactured and tested experimentally. Fig. 2 shows the photograph of the developed X-band prototype antenna including the SMA connector used to interface the antenna to the test equipment. As can be seen from the photograph, the ground-planes of the CPW transmission are connected via the SMA adaptor.

### III. Results

Fig. 3 shows the simulated and measured return loss of the proposed coplanar waveguide-fed quasi-Yagi antenna. As can be seen from the plot, the antenna operates from 7.7 to 12 GHz covering the required X-band. The simulated result closely resembles the measured result at the lower and upper resonant frequency validating the design of the antenna. However, slight
A CPW-fed antenna has been presented. The antenna is one of the simplest forms of a planar quasi-Yagi and it does not require any complicated balun structure and is also uniplanar. The 10 dB return loss bandwidth of the antenna is 44% and the measured gain varies between 3.4–7.4 dBi across the impedance matched bandwidth. The front-to-back ratio of the antenna was measured as 15 dB. The antenna is small in size indicating that it is a good candidate for phased arrays.

REFERENCES

Compact broadband coplanar waveguide-fed curved quasi-Yagi antenna

H.K. Kan, A.M. Abbosh, R.B. Waterhouse and M.E. Bialkowski

Abstract: A novel uniplanar coplanar waveguide-fed quasi-Yagi antenna is presented. The innovativeness of this design is because of its feed selection and its elliptical structure and in doing so achieving a wide bandwidth and compact size. An X-band prototype is developed and measures a bandwidth of 40%, with 3.2 dBi gain and 11 dB front-to-back ratio measured at 10 GHz. The antenna is realised on a high dielectric constant substrate and is compatible with monolithic microwave integrated circuits and solid-state devices.

1 Introduction

There are many applications in communication systems that necessitate the use of antenna arrays. These include phased arrays, power combing arrays and multi-beam communication arrays [1–3]. Planar quasi-Yagi is a promising candidate because of its several advantages such as low profile, low cost, lightweight and ease of integration into planar arrays, highly compatible with solid-state devices and monolithic microwave integrated circuits (MMICs). Various designs of the planar quasi-Yagi antennas have been reported in the literature covering the X-band of the frequency spectrum [4–13]. In each of the design, the driven and the driver elements are similar; however, the most evident distinction is the feeding mechanism used. Typical feeding technique frequently employed includes the microstrip feed connected to a broadband microstrip-to-coplanar stripline (CPS) [1, 4, 5] and the coplanar waveguide-fed (CPW) requiring a balun to transform the CPW mode at the input port to the CPS [6]. The effect of the balun on the quasi-Yagi antenna has been investigated [7]. Each has its own merits; however, the major drawback is its complicated feed requiring balun. A simplified feed and modified printed Yagi was reported recently to alleviate the design complexity with the baluns [8]. In this design, the feeding structure consists of a microstrip line that transitions to a transmission line formed by two parallel strips. One side of the driver element is connected to the feed whereas the other side of the driver element is connected to the ground connected via a CPS. Although this feeding method and driver configuration do not require additional devices such as balun or microstrip-to-CPS balun, it needs two sides of the substrate to be etched. This, in fact, reduces the space where active devices and MMIC components can be placed.

We present a single-layer CPW quasi-Yagi antenna. The antenna utilises the CPW feed, therefore alleviating the design complexity commonly associated with the feeding network of the quasi-Yagi antenna. In addition, the driven element, directors and the ground have curved structure created by elliptical shapes resulting in a small and compact Yagi. The measured 10 dB return loss bandwidth of the antenna is 40% covering the X-band. The antenna is compact where the overall dimension of the antenna including the substrate is 0.3λ0 × 0.5λ0 (where λ0 is the free space wavelength at the design frequency).

2 Configuration and design strategy

The schematic of the proposed antenna is shown in Fig. 1. As can be seen from the figure, the antenna consists of two director elements, a driven element and a ground plane acting as a reflector. The driven and director elements are created by having ellipses of radii r1 and r2 and its ratio of r1 and r2, respectively. The director elements are located from the centre of the driven element separated by distance of Sdir. The antenna is CPW reducing the complexity commonly associated with having broadband baluns for matching the driven element and the antenna feed. The compactness of this design is because of its elliptical elements and tapered ground plane where the currents of the antenna are extended resulting in it achieving a smaller volumetric size compared to straight structures [14]. As a starting point to design the antenna, the length of the driven element should be around 0.5λeff whereas the lengths of the directors should be in the order of 0.45λeff according to the Yagi design principles [15]. Here, λeff refers to the effective wavelength at the lowest frequency of operation, in this case, at 8 GHz. It is calculated assuming the following value for the effective dielectric constant of the substrate, εr = (εr + 1)/2, where εr is the dielectric constant of the substrate. The distance between the directors should be between 0.1 and 0.2λeff. To develop a compact antenna, we chose the value 0.14λeff. The design frequency for the proposed antenna is 8 GHz which results in the following initial values of the antenna parameters assuming the substrate to be Rogers RT6010 (0.64 mm, εr = 10.2): Ldir = 8 mm, Ld1 = Ld2 = 7.2 mm, Sdir ≈ Sd1 ≈ 1.6 mm. The commercial package, Ansoft HFSSv9.2, based on a 3D full-wave finite element method...
was then utilised in assisting to optimise the proposed antenna for wide bandwidth covering the X band. The antenna was optimised after simulation and the following dimensions were selected: 

- \( L_{dir} = 8 \text{ mm} \), 
- \( L_{dir1} = 7.3 \text{ mm} \), 
- \( L_{dir2} = 7.7 \text{ mm} \), 
- \( S_{ref} = 2.1 \text{ mm} \), 
- \( S_{dir1} = 1.71 \text{ mm} \), 
- \( S_{dir2} = 1.76 \text{ mm} \), 
- \( L = 12 \text{ mm} \), 
- \( W = 20 \text{ mm} \), 
- \( W_{dir1} = 1.3 \text{ mm} \), 
- \( W_{dir2} = W_{dir3} = 1.2 \text{ mm} \), 
- \( r_1 = 6 \text{ mm} \), 
- \( r_{a1} = 0.5 \), 
- \( r_2 = 4.82 \text{ mm} \), 
- \( r_{a2} = 0.38 \).

In this optimised design, the antenna operates from 8 to 12 GHz. The total area of the substrate is \( 0.3 \lambda_0 \times 0.5 \lambda_0 \). It is interesting to note that the ground plane has a significant effect on the bandwidth and resonant frequency of the antenna. As the ground plane is brought closer to the driven element (the distance \( S_{ref} \)), the resonant frequency is lower albeit at the cost of reduced bandwidth. It is interesting to note that the distance \( S_{ref} \) performs the task of a balun, therefore alleviating the need for the complicated balun commonly required. This simply implies that the bandwidth can be controlled by carefully optimising this variable. The current design features two director elements. Incorporating additional elements has the potential for increasing the gain or bandwidth of the antenna [5]. However, this also increases the number of design parameters as well as the complexity of the design optimisation, and has not been investigated extensively here but has been reported previously [16, 17] and these also includes the parametric study of the quasi-Yagi antenna. The use of tapered ground plane enables the current to have a smooth transition allowing the broadband nature of the antenna. The antenna is manufactured and tested experimentally. Fig. 2 shows the photograph of the developed X-band prototype antenna. As can be seen from the photograph, the antenna is etched on one side of the substrate enabling components to be placed on the other side of the substrate. The ground planes of the CPW transmissions are connected via the SMA coaxial cable. The bare substrate in front of the antenna is required to achieve a bandwidth capable of covering the whole of X band.

## 3 Results

The simulated and measured return loss performance of the proposed compact CPW-fed quasi-Yagi antenna is shown in Fig. 3. As can be seen from the plot, the antenna operates from 8 to 12 GHz covering the required X band. The simulated result closely resembles the
measured result at the lower and upper resonant frequency points validating the design of the proposed antenna. However, considerable discrepancy can be observed between the simulated and measured results for the ripples below −10 dB. This can be attributed to the connection between the feed point of the antenna and the cable used to connect the antenna to the SMA connector as shown in the photograph of Fig. 2 as this was not included in the simulation setup. The far-field radiation patterns were measured in an anechoic chamber at 8 and 10 GHz and show similar characteristics; however, for brevity, only the 10 GHz radiation patterns for the two principal planes, namely E- and H-plane co-polarised fields, are shown in Fig. 4. As can be seen from the figure, there is slight tilting of the beam and this is because of the little asymmetry of the antenna geometry. This can be alleviated by wire-bonding the ground planes [18]. It is worth noting that the E- and H-plane cross polarised fields were at least 11 dB lower than the co-polarised fields and show omni-directional radiation patterns, is not shown here. As can be seen from the plot in Fig. 4, there is directivity in the two principal planes (note that 90° is the endfire radiation direction on the plots) with the front-to-back ratio of 11 dB. Although this value is slightly lower than the previously reported quasi-Yagi [5, 6, 8, 9], this is anticipated given its compact size. According to the results presented in [15] for Yagi antennas, it is possible to increase the front-to-back ratio by increasing the distance between elements of the antenna to a certain limit. Fig. 5 shows the measured gain of the CPW-fed quasi-Yagi for frequencies from 8 to 12 GHz. The gain of the antenna varies from 2.25 to 3.2 dB over the whole band. The nominal radiation efficiency of the antenna estimated using Ansoft HFSSv9.2 is at least 90% across the impedance-matched band. The antenna was simulated in a coplanar and stacked array configuration using HFSS and it was found that the mutual coupling is less than 25 dB when the spacing between the antenna elements is $\lambda/2$ or larger making it suitable for array applications.

4 Conclusion

A CPW-fed quasi-Yagi antenna has been presented. The antenna is uniplanar and does not require any complicated balun structures. The 10 dB return loss bandwidth of the antenna is 40% and the measured gain varies between 2.25 and 3.2 dBi across the impedance-matched bandwidth. The front-to-back ratio of the antenna was measured as 11 dB. The antenna is compact in size indicating that it is a good candidate for phased arrays and can easily be integrated with active monolithic microwave integrated devices.

5 References

Design of Ultrawideband Planar Monopole Antennas of Circular and Elliptical Shape

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Abstract—An efficient approach is described for designing ultrawideband (UWB) antennas in the form of planar monopoles of elliptical and circular shape. To avoid the time consuming trial-and-error approach presented in other works, simple design formulas for this type of radiators are described and their validity is tested via electromagnetic simulation and measurements. Full electromagnetic wave investigations are performed assuming three types of substrates with wide range of dielectric constant and thickness. The presented results show that the proposed method can be applied directly to design planar antennas that cover the ultrawide frequency band from 3.1 GHz to more than 10.6 GHz. Four types of monopole antennas were manufactured using RT6010LM substrate and their operation was tested in terms of return loss, radiation pattern characteristics, gain, and time domain response. The developed antennas feature UWB behavior with near omnidirectional characteristics and good radiation efficiency. The time domain transmission tests between two identical elements show that the manufactured circular antennas offer better performance in terms of distortionless pulse transmission than their elliptically shaped counter parts. These antennas are also assessed in terms of fidelity factor. The manufactured antennas show a high fidelity factor which is more than 90% for the face-to-face orientation.

Index Terms—Antenna design, planar antenna, ultrawideband (UWB) antenna.

I. INTRODUCTION

ANY EMERGING microwave techniques and applications aim at using ultrashort pulses on the order of nanoseconds. In the frequency domain, such signals occupy an ultrawideband (UWB) frequency spectrum. In 2002, US-FCC has assigned the frequency band of 3.1–10.6 GHz with respect to these emerging UWB activities [1]. The primary objective of UWB is the possibility of achieving high data rate communication in the presence of existing wireless communication standards. For example, the recent IEEE protocol 802.11 g provides only 54 Mbps data rate. The use of UWB can give data rates of the order of hundreds of megabits per second. In addition to wireless communications, the use of UWB signals is envisaged in microwave imaging applications. This is motivated by the fact that such signals offer an increased resolution of imaged objects [2]–[4].

Radio and imaging systems, employing UWB, require suitable antennas as transducers between UWB transceivers and the propagating medium. To this purpose, several planar monopole antennas with various shapes have been devised [5]–[14]. The shortcoming of these planar UWB antenna designs is that they are based on the lengthy trial and error method that involves computationally intensive full wave electromagnetic simulations. When one decides to design an antenna using a different dielectric substrate, the time consuming design process has to be fully repeated. In such circumstances, the designers are interested in having simple design formulas that provide a very good approximation to the final design when sophisticated EM analysis and design software packages are applied. The present paper addresses this issue and provides simple design formulas, which are suitable for UWB antennas in the form of planar monopoles fed from a microstrip line. It is shown that the difference between values of the design parameters, i.e., dimensions of antenna structure, is less than 10% compared with the optimized values obtained using the commercial software Ansoft HFSS [15]. The paper is organized as follows. Section II describes the proposed design method. Section III presents results of simulations using Ansoft HFSS. Section IV reports on experimental results and Section V concludes the findings of this paper.

II. DESIGN

Configurations of the UWB antennas, which are investigated here, are presented in Fig. 1. The structures shown in Fig. 1(a) and (c) are created by a planar conducting surface formed by the intersection of either two ellipses or two circles in a two-side conductor-coated substrate. The primary radiating element and the microstrip feed are on one side of substrate while the ground plane is on its other side. In these structures, the surface electric current flowing on an elliptical or circular shaped conductor can be regarded as the primary source of radiation. Here, we call them E-monopoles. In turn, the structures shown in Fig. 1(b) and (d) are complementary to those in Fig. 1(a) and (c) and are named here as planar EC-monopoles.

The use of the terms of planar monopoles requires an extra explanation. This is because the considered antennas resemble other types of antennas known in the antenna literature. The initial structure of planar monopole formed by a square patch vertically positioned above a horizontal ground plane and fed from a coaxial line was introduced in [16]. The extension of this concept was made in [8], where the patch and the finite size ground were proposed to be formed in one plane. The justification for the use of the name “monopole” in [16] stemmed from the fact that a coaxially fed wire monopole was stretched and...
The feeder, are calculated [18] as being half and quarter of the effective wavelength \( \lambda_{ef} \) at the frequency \( f \). Therefore \( \lambda_{ef} = c \sqrt{\varepsilon_{\text{eff}}} f \) is \( \lambda_{ef} = \frac{\lambda_{ef}}{2} \) and \( \lambda_{ef} = \frac{\lambda_{ef}}{4} \), where \( c \) is speed of light in free space.

The ground plane of the elliptical (circular) monopoles is in the shape of a half ellipse (circle) whose dimensions are chosen to be similar to those for the larger ellipse (circle) of the radiating structure. This choice is made in order to get a smooth tapering between the radiating structure and the ground plane.

Centers of the two ellipses or the two circles are chosen such that the width of the radiator at the feeding point is equal to width of the microstrip feeder, whereas width of the slot \( \rho \) between the radiator and the ground plane is around half of the feeder width. Therefore \( f_1 = \frac{f_1}{2} \), \( f_2 = f_2 + \frac{f_2}{2} \), for the E-monopoles and \( f_2 = f_2 \) for the EC-monopoles, where \( f_1 \) and \( f_2 \) are centers of the large and small ellipses (or circles) measured from the end of the feeder.

As noted earlier in this section, each of the structures shown in Fig. 1 can be viewed as a vertical monopole located a quarter of the effective wavelength above a finite ground plane or a horizontal bow-tie dipole of approximately half-effective wavelength in the horizontal (\( \phi \)) and vertical (\( \theta \)) directions. All of these structures include a smooth tapered slot between the monopole and the ground. Its opening is about half of the effective wavelength at the lowest frequency of operation.

III. RESULTS OF SIMULATIONS

The presented design method uses the effective dielectric constant \( \varepsilon_{\text{eff}} \), given as the average value of the relative permittivity of the two mediums (substrate and air), to work out the dimensions of the E and EC monopoles. In practice, one can expect that this effective dielectric constant can also depend on the substrate thickness and the operational frequency. Therefore the formula \( \varepsilon_{\text{eff}} = \left( \varepsilon_r + 1 \right)/2 \) can give only the first order approximation to work out optimal dimensions of the E and EC monopoles. In order to investigate this approximation, we undertake full electromagnetic wave investigations for a range of substrates of various permittivity and thickness.

First, we present a number of designs assuming DuPont951 material with \( \varepsilon_r = 7.8 \), length \( \text{length} = \{10, 15, 20\} \) and thickness \( \text{thickness} = 1 \) mil. Parameters for UWB elliptical and circular E and EC monopole antennas obtained using the above presented design formulas are shown in Table I assuming a 2.5 GHz value for \( f_1 \). Table I also includes the optimum values of the design parameters obtained using Ansoft HFSSv10. This optimization process aimed at near omnidirectional characteristics of the antennas with 10 dB return loss bandwidth of at least 3.1 to 10.6 GHz. A comparison between the dimensions obtained from the use of the proposed formulas and those generated using Ansoft HFSS is shown in Table I. The difference is less than 10\%, and thus indicates that the proposed design formulas are quite accurate. The presented values show that the designed UWB antennas are of compact size.
size (in comparison with the operational wavelength), which is advantageous in applications requiring compact RF front ends.

Fig. 2 shows variations of return losses with frequency for the four designed monopole antennas assuming DuPont951. The obtained results indicate that all of the designed antennas have UWB characteristics with an impedance bandwidth covering at least 3.1–10.6 GHz assuming a 10 dB return loss reference. As seen in Fig. 2, the 10 dB return loss bandwidth for the E type elliptical and circular monopoles commences at a frequency (of about 2.5 GHz) lower than the EC-type counterparts (approximately 3 GHz).

From the UWB applications point of view, the antenna is usually required to have an omnidirectional radiation pattern (here in the $\varphi$-plane). This requirement is well fulfilled in the lower part of UWB (3 and 5 GHz), where almost a perfect donut-shape radiation pattern is observed, as shown in Fig. 3 for the elliptical E-monopole. However, it slowly diverges from ideal at the upper frequencies (7 GHz and above). This comes from the fact that the tapered slot between the monopole and the finite ground, being part of this antenna, is responsible for forming a directional pattern in the $\varphi$-plane. As a result, we obtain a wide beam in the direction along the slot while shallow nulls are observed in the directions orthogonal to the slot. Similar results were obtained for the other designed antennas. They are confirmed by measurements in Section IV.

Gain of the designed antennas is revealed in Fig. 4. It varies between about 0.5 to 4.7 dB over the required bandwidth for the E-monopoles, whereas it is between −10.5 to 2.5 dB for the EC-monopoles. These results indicate that the E-monopoles show a higher gain compared with their complementary counterparts (EC-monopoles).

Using a substrate with a high dielectric constant and a direct microstrip feeder may cause deterioration in the radiation efficiency of the antenna (meant as the ratio of power radiated to power delivered to the feeder). To check this case, variation of the radiation efficiency with frequency for the designed antennas was calculated with HFSS software. These calculations

### Table I

<table>
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<th>Parameter</th>
<th>Calculated Values</th>
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<th>Optimized Values</th>
<th>Optimized Values</th>
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Fig. 2. Variation of return loss with frequency for the designed antennas.

Fig. 3. Three dimensional radiation pattern for the elliptical electric monopole at different frequencies.

Fig. 4. Variation of the peak realized gain with frequency for the designed antennas.
have shown that the designed antennas have always a good efficiency which is greater than 92%.

In order to test the validity of the design method for different types of substrates with wide range of thickness (h), it was applied to design elliptical E-monopoles using the following substrates: Rogers RO4003C with \( \varepsilon_r = 3.58 \) and substrate loss = 0.012%. Rogers RT6010LM with \( \varepsilon_r = 1.12 \) and substrate loss = 0.012% in addition to DuPont951. The thickness of the above three substrates was varied from 0.5 to 1.5 mm. The software HFSSv10 was used to simulate performance of the antennas with the dimensions shown in Table II. No optimization of dimensions was used in this case. The results of the simulation for the return loss are shown in Figs. 5–7. It is apparent that the effect of the substrate thickness on the return loss behavior is small with respect to the lower frequency range of the UWB, which is from about 3 to 8 GHz. This is especially true for low dielectric constant substrates (the case of \( \varepsilon_r = 3.58 \)). The variation in return loss as a function of substrate thickness becomes more pronounced for frequencies above 8 GHz and for substrates with a larger relative dielectric constant (\( \varepsilon_r = 7.8 \) and \( \varepsilon_r = 1.12 \)). However, irrespective from the substrate thickness and its permittivity, the presented formulas deliver monopoles dimensions which cover the required UWB 10 dB return loss bandwidth, with exception of \( \varepsilon_r = 3.58 \) and \( h = 1 \) mm where the return loss becomes slightly lower than 10 dB for frequencies between 8 and 10.6 GHz.

The design steps for the antennas were repeated assuming this time different values for the lowest frequency of operation (\( f_1 \)). The substrate assumed in this simulation was DuPont951 with

<table>
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</table>

Fig. 5. Variation of the return loss with frequency for different values of substrate thickness. The substrate is RO4003C with \( \varepsilon_r = 3.38 \).

Fig. 6. Variation of the return loss with frequency for different values of substrate thickness. The substrate is DuPont951 with \( \varepsilon_r = 7.8 \).

Fig. 7. Variation of the return loss with frequency for different values of substrate thickness. The substrate is RT6010LM with \( \varepsilon_r = 7.8 \).
Fig. 8. Comparison between the designed low frequency and the results obtained via simulation. The substrate used is DuPont951 with $\varepsilon_r = 7.8$ and $h = 0.3$ mm.

Fig. 9. (a) Photographs of the manufactured elliptical E-monopole, (b) elliptical EC-monopole, (c) circular E-monopole, and (d) the circular EC-monopole.

0.5 mm thickness, and $f_1$ was considered to vary from 1.5 GHz up to 3 GHz. Simulation results for the return losses in the frequency band of 1 to 6 GHz are shown in Fig. 8. With respect to the commencement of the 10 dB return loss bandwidth, there is only a 5% difference between the assumed $f_r$ and the values shown in the plots in Fig. 8. This confirms that the presented design method can very well predict the commencement of the 10 dB return loss bandwidth for all of the investigated planar monopole antennas.

IV. EXPERIMENTAL RESULTS

The four types (circular and elliptical of E and EC type) of UWB printed monopole antennas were manufactured using the substrate Rogers RT6010LM with thickness 0.64 mm. The photographs of the manufactured antennas including coaxial feeding ports are shown in Fig. 9. The design parameters for the developed antennas were calculated using the proposed method and the values were: $r = 25$ mm, $l = 13$ mm, $t_1 = 25$ mm, $t_2 = 13$ mm, and $\alpha = 0.5$ mm.

The measured results for the return loss are presented in Fig. 10. Return loss measurements were obtained using HP8510/HP8530 network analyzer in an anechoic chamber. The results shown in Fig. 10 indicate that the four types of antennas feature UWB behavior with bandwidth from 3.1 GHz to more than 15 GHz assuming a 10 dB return loss reference.

The measured far field radiation patterns of the E-monopole antenna in the two principal planes, $\phi$; plane ($\phi = \Phi$) and the $\theta$; plane ($\phi = 90^\circ$) are shown in Fig. 11 at different frequencies. The radiation patterns reveal a near omnidirectional behavior in the $\phi$; plane. A better quality omnidirectional pattern in the $\phi$; plane is observed at lower frequencies (3–6 GHz) of UWB. This agrees well with the finding obtained from the computer simulated radiation patterns, as reported in Section III. The measured radiation patterns of the other antennas exhibit almost the same behavior and therefore they are not shown here. In addition to radiation patterns, the antenna gain was also measured. The measurement was performed using a standard double ridge corrugated horn antenna as a reference gain antenna. The distance between the transmitter and the receiver was 1 m. The results of measurements presented in Fig. 12 show that the gain is between 0.5 and 3.9 dB for the E-monopoles and between 0.1 and 2.5 dB for the EC-monopoles in the frequency range 3 to 10 GHz. Radiation efficiency of the four antennas was calculated using Ansoft HFSSv10 and it was found to be higher than 90% across the whole band.

The last test concerned the ability of the manufactured antennas to transmit and receive pulses without distortions. In this
that changing the orientation of the receiving antenna has only a little effect (more pronounced for E-monopoles) on the impulse response. This confirms the omnidirectional behavior of the developed antennas.

The last step in the impulse response measurement is to calculate the fidelity factor, which in turn is related to the results shown in Fig. 13. Using the method presented in [19], it was found that the manufactured antennas had more than 90% fidelity factor in case of the circular antennas and around 80% in case of elliptical antennas, for any orientation. Similar to the results shown in Fig. 13, the best fidelity results were obtained for the monopoles facing each other (0-degree orientation case). In this case, all the manufactured antennas offered more than 90% fidelity. These values of fidelity factor as well as the small amplitude distortions observed in Fig. 13 confirm the high capability of the developed antennas to send and receive UWB pulses with only small distortions.

V. CONCLUSION

In this paper, a simple method for designing compact UWB planar monopole antennas of elliptical and circular shape has been presented. For the chosen configurations, the radiating elements are formed by the intersection of two ellipses or two circles. An explanation and justification for the use of the term planar monopole for these structures has been given. Full electromagnetic wave simulations have shown that the proposed design method is valid for a wide range of dielectric constants and substrate thickness. The four proposed types of antennas have been manufactured using RT6010LM substrate. The measurements have shown that the proposed design formulas enable the development of UWB antennas with suitable radiation characteristics. The designed antennas cover the 3.1–10.6 GHz band allocated to UWB systems in terms of return loss performance with well behaved omnidirectional radiation pattern and more than 90% radiation efficiency. The time domain test of transmission between two identical antennas at different orientations has shown a better performance of the circular shaped monopoles. This has been confirmed by calculations of the fidelity factor being more than 90% for the face-to-face orientation.

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REFERENCES


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Design of UWB Planar Antenna With Improved Cut-Off at the Out-of-Band Frequencies

Marek E. Bialkowski and Amin M. Abbosh

Abstract—This paper describes a method to improve the cutoff capability of an ultra wideband (UWB) planar antenna at the out-of-band frequencies using a meandered slot. In the presented UWB design, aimed for operation between 3.1 and 10.6 GHz, the antenna is formed by a planar monopole and a ground plane both of half circle shape, with a meandered-shape slot made in the monopole. In order to assess the antenna’s cutoff capability for the out-of-band frequency range (below 3.1 GHz and above 10.6 GHz), two antennas, without and with a slot, are designed and developed. The simulated and measured results show that the meandered slot improves the cutoff capability of the antenna by decreasing the return loss in the 2–3 GHz band from 6–10 to 1 dB, and in the 10.6–11 GHz from 11 to 5.5 dB. At the same time the antenna’s performance in the passband is unaffected both in terms of the return loss and radiation pattern. It is also shown that the antenna is useful for the pulsed UWB applications as the measured impulse response of the antenna indicates a distortionless pulse transmission.

Index Terms—Planar antenna, ultra wideband (UWB) antenna.

I. INTRODUCTION

ULTRA Wideband (UWB) technology has gained a lot of popularity among researchers and the wireless industry after the FCC permitted its marketing within the frequency band of 3.1 to 10.6 GHz [1]. With respect to wireless communications, its use is aimed at obtaining high capacity short-range links with low-cost low-energy transceivers. To establish the communication between two nodes, the transceivers require UWB antennas, preferably of small size and low manufacturing cost. Various shape planar monopole antennas with coaxial, microstrip, or coplanar waveguide feeds have been proposed as candidates to fulfill this requirement.

The design of ultra wideband antennas has been a hot research topic, especially during the last few years. In order to improve coexistence of UWB with other wireless standards, a considerable amount of research has been devoted to devising techniques to reject certain bands within the passband of the UWB [1]–[5]. It has to be noted that most of the presented works concern the rejection of a single band within the passband of UWB [1]–[5]. The presented design relies on the use of a meandered slot which is made in the radiating planar monopole. In order to generate an efficient frequency cutoff, the slot requires a certain length, which is calculated using a simple formula. Also, it requires a suitable shape. This is to accommodate it within a limited area of the monopole and to prevent rejecting frequencies within the passband of UWB. The effectiveness of the proposed method is demonstrated via full-wave simulations and measurements on two UWB antennas: one without a slot and the other with a slot.

II. DESIGN

The proposed configuration of the planar UWB antenna with a sharp cutoff at the low frequency band is illustrated in Fig. 1. In this antenna, both the radiating element and a ground plane structure are in the form of half circle. A meandered slot is formed in the radiating element. The antenna is fed using a 50-Ω microstrip line whose width is calculated using the well-known microstrip line design equations [6].

Fig. 1. Configuration of the UWB antenna with the meandered slot.

The presented design relies on the use of a meandered slot which is made in the radiating planar monopole. In order to generate an efficient frequency cutoff, the slot requires a certain length, which is calculated using a simple formula. Also, it requires a suitable shape. This is to accommodate it within a limited area of the monopole and to prevent rejecting frequencies within the passband of UWB. The effectiveness of the proposed method is demonstrated via full-wave simulations and measurements on two UWB antennas: one without a slot and the other with a slot.

\[ l = \frac{c}{2\sqrt{\left(\frac{f_1}{f_2}\right) + \frac{1}{2}}} \]

where \( c \) is speed of light. Note that the length and width of the antenna structure, according to (1), are equal to half of the effective wavelength at the lower frequency \( f_1 \).

As the radiating element and the ground are separated, the design also requires a suitable choice of the spacing \( s \). A parametric analysis (using the full-wave electromagnetic simulator Ansoft HFSS v10) concerning the choice of the best value for \( s \) addresses this uncared issue by designing a UWB antenna with sharp cutoffs at the low frequency band (below 3.1 GHz) and the high frequency band (above 10.6 GHz).

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indicates that it should not be larger than thickness of the substrate \(t_0\) in order to get the widest operational bandwidth of the chosen UWB antenna configuration.

The above simple procedure results in an antenna design which covers the required UWB range from 3.1 to 10.6 GHz. This is with respect to a 10-dB return loss used as reference. The next step concerns the rejection of frequencies below 3.1 GHz and above 10.6 GHz, as recommended by FCC.

In order to improve the cutoff properties of the designed antenna at the out-of-band frequency ranges, we propose the use of a meandered-shape slot with a total length \(l_s\) made in the radiator. This is shown in Fig. 1. It has to be noted that the use of a slot in a UWB planar monopole to reject the middle bands (4.9–5.9 GHz) has already been demonstrated by many researchers. The works shown in [2] and [3] are typical examples. However, these solutions are unsuitable for obtaining a sharp cutoff at 3.1 GHz. The reasons are the shape, size, and location of the slots relative to size and shape of the radiator. The proposed straight or circular shapes in [2] and [3] do not allow for accommodating a slot to reject the frequencies below 3.1 GHz. Another issue that has been omitted in previous considerations is that such slots can also cause rejection at harmonic frequencies, which are integer multiples of the fundamental frequency that is rejected.

The outlined problems are overcome in this letter by a meandered slot located very close to the feeding element of the UWB antenna.

Similarly as in the previous designs [2], the total length of the slot is chosen to be equal to half of the effective wavelength, however, this time it is calculated at frequency \(f_c\), which represents centre of the low out-of-band frequency range. For the UWB antenna under consideration, the frequency \(f_c\) is chosen to be 2.3 GHz. Therefore, the slot’s total length \(l_s\) is selected according to the following expression:

\[
l_s = \frac{c}{2f_c\sqrt{(\varepsilon_r+1)/2}}.
\]

With this choice of the slot length, the slot improves the rejection capability of the antenna at the fundamental frequency \(f_c\). Potentially, this slot length is also responsible for rejecting harmonic frequencies of \(f_c\) (\(f_c = nf_{10}\), where \(n\) is a positive integer number). Assuming that the slot is symmetric with respect to the monopole, its feed and ground plane, only odd harmonics need to be considered. In order to improve the rejection at the high out-of-band frequency (above 10.6 GHz) the fifth harmonic of \(f_c\) needs to be used. At the same time, the third harmonic of \(f_c\) has to be eliminated. This is because it can cause the rejection within the passband of the UWB antenna. To eliminate effect of the third harmonic, the shape of the slot has to be chosen such that there is a destructive coupling between different sections of the slot at the third harmonic.

### III. Results

The design of the proposed UWB antenna is undertaken assuming Rogers RO4003C substrate with a dielectric constant equal to 3.38, tangent loss \(\tan\delta = 0.0027\), and thickness \(t_0 = 0.508\) mm. The design makes use of the presented formulas followed by the optimization with the software HFSSv10.

The first design concerns a UWB antenna without a slot. Values of the design parameters shown in Fig. 1, which were calculated using the presented method and then optimized using HFSSv10, are: \(\varepsilon_r = 2.24\), \(l_1 = 23.5\) mm, \(\varepsilon_f = 1.18\) mm, \(s = 0.45\) mm. Concerning width and position of the tapered slot between the radiator and the ground, the optimization results of HFSS indicated that the best performance can be obtained when \(\varepsilon_s = 0.6\) mm, and \(\rho_s = 2\) mm.

The next design concerns a UWB antenna with a sharp rejection at \(f < 3.1\) GHz. In this case, a symmetric meandered-shape slot is made in the radiator of the UWB antenna, which was designed in the previous step. Length of the meandered slot is calculated using (2) after assuming the rejected band centred at 2.3 GHz. The calculated and optimized total lengths of the slot are equal to 44 mm, and 40.3 mm, respectively. It is clear that the presented method gives an accurate estimation of the required parasitic lengths where the difference between the calculated and the optimized values is about 8%. The chosen shape to reduce the coupling of the third harmonic and maintain the coupling of the fifth harmonic is shown in Fig. 1. The shape of this slot was chosen with the help of the parametric analysis capability of the software HFSS.

In the next step, the two antennas (without and with the meandered slot) were manufactured and experimentally tested. Fig. 2 shows variation of the simulated and measured return loss with frequency for the developed antennas. The return loss of the antenna without the slot reveals UWB behaviour with bandwidth from 3.1 GHz to more than 11 GHz assuming the 10 dB level as a reference. The return loss at the frequency range below 3.1 GHz varies slowly between 9 dB at 3 GHz to 4.3 dB at 2 GHz, while it is about 11 dB at the high out-of-band frequency range (above 10.6 GHz). This means that the antenna can still efficiently radiate in those out-of-band frequency ranges. Such undesired radiation can easily occur when a UWB link employs UWB pulses without limiting their spectrum with UWB bandpass filters.
Concerning the antennas with a slot, Fig. 2 shows that the rejection at the low frequency band is improved, where the return loss reduces sharply from 10 dB at 3.1 GHz to around 1 dB at 2.2 GHz. Over the passband (3–10.6 GHz) the behavior of the antenna is maintained. It can be also noticed from Fig. 2 that the antenna with a slot improves cut-off at the high frequency range, i.e., at $f > 10.6$ GHz. The measured return loss for the antenna without a slot is equal to 11 dB at 11 GHz, which means unwanted efficient radiation at this frequency, whereas it is 5.5 dB at 11 GHz for the antenna with a slot. It can be seen from the results of Fig. 2 that the effect of third harmonic, which appears at around 6.3 GHz, is considerably reduced to almost a negligible value. These results confirm that the chosen slot length and shape offer rejection of the fundamental and the fifth harmonic, while the third harmonic is passed almost without any attenuation.

From the UWB applications point of view, the UWB antennas are usually required to have an omnidirectional radiation in the plane orthogonal to the radiating element. Concerning the designed antennas, this requirement is fulfilled over the whole passband of 3.1–10.6 GHz, as shown in the measured results in Fig. 3, which depicts the radiation pattern at different frequencies for the two principle planes ($\phi_r$-plane and $\phi_s$-plane) for the antenna with a slot. With respect to these results, the antenna is in the $\phi_r$-plane with the width $r$ extending along the $r$-axis. Similar radiation patterns were observed for the antenna without a slot as indicated in Fig. 3(b) and (c).

In pulsed UWB applications, the time domain response of the antenna is an important parameter as it is required to transmit/receive very short pulses without distortion. Concerning the manufactured antennas, the measured impulse response indicates almost distortionless transmission as depicted in Fig. 4, where small distortions are present at a level less than 0.1 with respect to the pulse peak value.

IV. CONCLUSION

In this paper, a meandered-shape slot made in the radiating structure has been shown to be an effective means to improve the cutoff performance of an ultra wideband planar monopole at its out-of-band frequencies. The design strategy has been demonstrated for a half circle planar monopole having a half circle shaped ground plane. Two samples of antennas have been designed and tested: one without a slot and the other with a slot. It has been demonstrated via simulated and measured results that the use of the meandered slot improves the cutoff capability of the antenna at the low out-of-band frequency range (below 3.1 GHz) and at the high out-of-band frequency range (larger than 10.6 GHz). In turn, the antenna maintains its UWB performance both in terms of return loss and radiation pattern in the passband of 3.1 to 10.6 GHz. Concerning the pulsed UWB operation, the measured time domain characteristic of the antenna has revealed a high fidelity (almost distortionless) pulse transmission.

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Miniaturization of Planar Ultrawideband Antenna via Corrugation

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Abstract—In this letter, a procedure to design a miniaturized planar antenna of ultrawideband (UWB) performance is presented. The proposed method utilizes corrugated radiator and ground plane of elliptical shapes to design an omnidirectional antenna of compact size. The simulated and measured results show that using the proposed method results in a surface area reduction by more than 50% compared with the optimized noncorrugated structure, which is designed using the previously published procedure, while maintaining the other radiation characteristics of the antenna. The time domain response of the miniaturized antenna reveals a distortionless operation with more than 90% fidelity factor.

Index Terms—Miniaturization, planar antenna, ultrawideband (UWB) antenna.

I. INTRODUCTION

With the rapid growth in mobile communications, and the ever increasing demand for high data rate mobile systems, number of radios on mobile platforms has reached a point that the available space for the antennas has become a serious problem. Hence, miniaturized antennas have become an imperative research area for both the academia and the industry.

The fundamental limitations of small antennas were addressed a long time ago by several authors [1]–[3]. These studies show that for single resonant antennas, the smaller is the maximum dimension of an antenna, the lower is its bandwidth [4].

The challenge with miniaturization of antennas has been the tradeoff between the bandwidth and the reduction of the physical size of the antenna. One of the methods adopted in antenna miniaturization is to use substrates with high dielectric constant to reduce the antenna’s resonant frequency because of the inverse relation of permittivity and the resonant frequency [5]. However, bandwidth of the antenna is also inversely proportional to the permittivity; as a result, bandwidth reduction is inevitable in this approach with the physical size reduction. In another method, high permeability metamaterials were used to reduce the resonant frequency because of the inverse relation between them, in addition to maintaining the bandwidth because of the direct proportionality between the bandwidth and the substrate’s permeability [6]. Recently, there have been other approaches to develop compact radiators [7]–[10]. In [7], the current distribution over a sectorial loop antenna was studied and the regions of low electric current were removed to achieve a light and compact structure. However, the designed antenna still has a large size, needs a large ground plane, and has a non-planar structure, which prevents its use in many applications. Metamaterials that have negative parameters, such as a negative dielectric constant [8], [9], or a negative permeability [10], have been utilized for antenna miniaturization. Although the analyses in those papers show that the antenna’s size can be significantly reduced when using materials with such negative parameters, the designed antennas show a narrowband performance.

In several papers published recently [11], [12], it has been shown that certain design guidelines can be followed to build compact ultrawideband (UWB) antennas of planar structure. In this letter, a simple procedure is used in conjunction with those design guidelines to miniaturize an omnidirectional UWB planar antenna using corrugated radiator and ground plane. The results presented in this letter show that the proposed method reduces the antenna’s surface area by more than 50%, while maintaining the ultrawide bandwidth, efficiency, and radiation pattern.

II. DESIGN AND RESULTS

Structure of the utilized antenna in its original configuration is shown in Fig. 1(a). The overall dimension of the antenna (the length $L$ and the width $W$) is chosen initially according to the design guidelines of [12] to be equal to half of the effective wavelength calculated at the lowest frequency of operation (3.1 GHz). The radiator at the top layer and the ground plane at the bottom layer are in the form of half an ellipse with a major diameter equal to $W$. There is a slot ($\sigma$), which is chosen initially to be equal to the substrate’s thickness, between the radiator and the ground plane. The secondary diameter of the ellipses representing the radiator and the ground plane is equal to the major radius minus the slot value $\sigma$. Substrate of the antenna is Rogers RT6010 with dielectric constant $\varepsilon = 10.2$ and thickness $d = 0.64$ mm.

Before starting the miniaturization process, the structure designed according to the aforementioned procedure was tuned to cover the UWB with the smallest possible size. This was achieved using the optimization capability of the commercial software CST Microwave Studio. The final dimensions for the antenna are: $L = 18$ mm, $W = 19.5$ mm, $\varepsilon = 0.8$ mm, and $d = 0.5$ mm. The simulated return loss of the antenna, which is shown in Fig. 2, reveals an UWB performance, where the bandwidth extends from 3.1 to more than 12 GHz assuming the 10-DB return loss as a reference.
To compensate for effect of the size reduction on the low-frequency performance, the radiator and the ground plane are corrugated by cutting strips from them in the manner shown in Fig. 1(c). Depth of the corrugations is chosen to be less than quarter of the effective wavelength ($\lambda/4$) at the lowest frequency of operation (3.1 GHz) so that each slot presents an inductive reactance to the passing wave, and thus increases the electric length of the structure [13]. Therefore, the corrugated radiator with those short circuited slots resonates at a lower frequency, as compared with a noncorrugated structure of the same dimensions. This phenomenon is employed in this paper to improve the radiation performance of the trimmed antenna, especially at the low end of the frequency band. The corrugations should have a depth equal to $\lambda/2$ at a frequency which is just outside the required band, such as at 11 GHz, to make them behave as a rejecting filter [14]. This secures the rejection of the frequencies outside the UWB and prevents the interference with any wireless system working at that band.

Width of the slots ($W_{s1}$ and $W_{s2}$) and the distance between them were initially chosen such that the slant distance ($d_s$) is less than $\lambda/2$ calculated at the highest frequency of operation (10.6 GHz) [13]. The slots’ parameters were then optimized using Microwave Studio and the final values were found to be: $L_s = 3$ mm, $d_s = 4$ mm, $W_{s1} = 0.7$ mm, and $W_{s2} = 0.5$ mm.

Performance of the trimmed-corrugated antenna was simulated and the result for the return loss is shown in Fig. 2. Making slots at the radiator and the ground plane in the second miniaturization step shown in Fig. 1(c) almost restores the UWB coverage of the antenna as depicted in Fig. 2, where the 10-dB return loss bandwidth extends from 3.5 to 10.4 GHz. It is to be noted from Fig. 2 that the return loss performance of the miniaturized antenna is superior to that of the full size antenna across the band 8–10.4 GHz. It is also worthwhile to refer to the filtering effect of the corrugated structure. The slots behave as a rejecting filter when their depth approaches $\lambda/2$ and they completely reject the frequencies larger than 10.6 GHz, whereas this was not the case before corrugating the structure as shown in Fig. 2.

To verify accuracy of the proposed method, a prototype of the miniaturized antenna was developed and tested using a vector network analyzer in an anechoic chamber. The measured result for the return loss is shown in Fig. 2, where the bandwidth extends from 3.7 to 10.6 GHz. Comparing the simulated and measured results for the corrugated antenna shows a good agreement between them, though there is about 5% difference in value of the frequency at which the 10-dB return loss bandwidth starts.

In order to make sure that the miniaturization process does not significantly change the radiation characteristics of the antenna, the radiation pattern before and after miniaturization was measured and compared at the two principle planes ($\theta_1$ and $\theta_2$). The comparison showed that the radiation pattern does not change due to miniaturization. Because of the very slight difference between the radiation pattern of the full size and the miniaturized antenna, only pattern of the miniaturized structure is shown in Fig. 3 at three frequencies. It is obvious from this figure that the miniaturized antenna has an omnidirectional radiation.

As another step for the method’s validation, gain and radiation efficiency of the antenna before and after miniaturization

Fig. 1. Configuration of the UWB antenna (a) before miniaturization, (b) after the first step, and (c) after the second step of miniaturization.

Fig. 2. Variation of the return loss with frequency before and after miniaturization.
Fig. 3. The measured radiation pattern of the miniaturized antenna.

were compared. Concerning the gain, the measured results depicted in Fig. 4 show that its variation is typical for the omnidirectional performance and the maximum gain is below 2 dBi. The general shape of the gain’s variation is almost the same for pre- and postminiaturization, although there is a gain reduction of about 0.7 dB due to miniaturization. Concerning the radiation efficiency, the calculated values using the software CST Microwave Studio, which are not shown here, revealed that the miniaturization has negligible effect on the efficiency, which was found to be more than 90% before and after miniaturization across the whole UWB.

In the last step of verification, the time domain response of the miniaturized antenna was measured. For this purpose, two antennas were put side by side at a distance of 30 cm between them. An UWB pulse synthesized in the vector network analyzer to cover the band 3.1–10.6 GHz was transmitted from the first antenna, and the received pulse by the second antenna was measured. The result is shown in Fig. 5, where amplitudes of the transmitted and received pulses were normalized to have a peak equal to 1. It is clear from Fig. 5 that the miniaturized antenna is efficient in the UWB pulse operation where the received pulse has a low distortion and the fidelity factor calculated using the measured response was found to be more than 90%. The distortionless time domain performance of the miniaturized antenna was also confirmed by the measured group delay shown...
in Fig. 6, which reveals less than 1-ns fluctuations in the group delay across the UWB.

After inspecting performance of the antenna before and after miniaturization, it is possible to conclude that the proposed miniaturization approach results in antenna’s surface area reduction of more than 50% (from surface area of 18 mm × 19.5 mm to 10.4 mm × 16 mm) with almost the same useful bandwidth as compared with the full-size antenna. The corrugated structure also enables the UWB antenna to reject the out-of-band frequencies, and thus restricts its bandwidth to the permitted range (3.1–10.6 GHz).

III. CONCLUSION

A method has been presented to design a miniaturized planar omnidirectional antenna with UWB performance. The proposed method utilizes corrugated radiator and ground plane of elliptical shapes. The simulated and measured results have shown that the proposed method reduces the required surface area of the antenna by more than 50% compared with the optimized structure designed using the previously published procedure, while maintaining the antenna’s other characteristics, such as the gain, efficiency, and radiation pattern. The time domain response of the miniaturized antenna has revealed a distortionless performance.

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Research Article

Directive Antenna for Ultrawideband Medical Imaging Systems

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A compact and directive ultrawideband antenna is presented in this paper. The antenna is in the form of an antipodal tapered slot with resistive layers to improve its directivity and to reduce its backward radiation. The antenna operates over the frequency band from 3.1 GHz to more than 10.6 GHz. It features a directive radiation with a peak gain which is between 4 dBi and 11 dBi in the specified band. The time domain performance of the antenna shows negligible distortion. This makes it suitable for the imaging systems which require a very short pulse for transmission/reception. The effect of the multilayer human body on the performance of the antenna is also studied. The breast model is used for this purpose. It is shown that the antenna has more than 90% fidelity factor when it works in free space, whereas the fidelity factor decreases as the signal propagates inside the human body. However, even inside the human body, the fidelity factor is still larger than 70% revealing the possibility of using the proposed antenna in biomedical imaging systems.

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1. INTRODUCTION

Ultrawideband (UWB) (3.1–10.6 GHz) microwave imaging is a promising method for biomedical applications such as cancer detection because of their good penetration and resolution characteristics. The underlying principle of UWB cancer detection is a significant contrast in dielectric properties, which is estimated to be greater than 2 : 1 between normal and cancerous tissue. UWB imaging systems have shown encouraging results in the detection of tumors for early breast-cancer detection [1].

In the UWB imaging systems, a very narrow pulse is transmitted from a UWB antenna to penetrate the body. As the pulse propagates through the various tissues, reflections and scattering occur at the interfaces. A particular interest is in the scattered signal from a small size-tissue representing a tumor. The reflected and scattered signals can be received using an UWB antenna, or array of antennas, and used to map different layers of the body. For an accurate imaging system with high resolution and dynamic range, the transmitting/receiving UWB antenna should be planar, compact in size, and directive with high-radiation efficiency and distortionless pulse transmission/reception.

The majority of the compact UWB antennas presented in the literature exhibit omnidirectional radiation patterns with relatively low gain and an impulse response with observable distortion [2]. These types of UWB antennas are suitable for the short-range indoor and outdoor communication. However, for radar systems, such as an UWB microwave imaging system for detection of tumor in woman's breast, a moderate gain directional antenna is advantageous. In addition to an UWB impedance bandwidth, as defined by the minimum return loss of the 10 dB, the UWB antenna is required to support a very short pulse transmission with negligible distortion. This is necessary to achieve precision imaging without ghost targets. The unipolar and antipodal Vivaldi antennas presented in the literature [3–5] satisfy the requirements for imaging systems in terms of bandwidth, gain, and impulse response. However, the achieved performance is at the expense of a significant size, which has a length of several wavelengths. Therefore, the challenge is to reduce their physical dimensions such that it can be incorporated in a compact microwave imaging detection system, while maintaining its broadband, high-gain, and distortionless performance.

Several UWB antenna designs with compact size and low distortion have been proposed for the use in the medical
imaging systems [6–8]. Each has its own merits and drawbacks. Some of the proposed antennas have a nonplanar structure, whereas others have low-gain and/or low-radiation efficiency. The low-radiation efficiency is a major impairment that limits the dynamic range of the imaging system, whose major objective is to detect a weak backscatter impairment that limits the dynamic range of the imaging system.

In the presented work, a compact (5 cm × 5 cm) elliptical tapered slot UWB antenna is described. A clear design guideline is given in order to show how to calculate values of the different design parameters of the antenna. Resistive layers were incorporated with the radiating elements of the antenna to improve its directivity and reduce any backward radiation which may affect the accuracy of the imaging system. The measured and simulated results of the proposed antenna show an ultrawideband behavior with a moderate gain and distortionless pulse transmission/reception.

2. DESIGN

The antenna presented in this paper is to be used in a microwave imaging for breast-cancer detection. The imaging system includes a circular array of the proposed ultrawideband antenna. In this system, one of the antennas is used to transmit a microwave signal while the rest of the antennas in the array receive the scattered signal. The measured data is collected and then the measurement procedure is repeated with the second transmitting the signal while the remaining are used for receiving the scattered signal. This process is repeated until all antennas in the array perform the transmitting role. The antenna array can be moved up and down automatically via a computer-controlled high-precision linear actuator. This facilitates the collection of multiple planar data for 3D object imaging.

The proposed ultrawideband antenna for inclusions in the UWB microwave imaging system is shown in Figure 1. It resembles an antipodal tapered slot antenna fed by a parallel strip line.

The radiating element is in the form of an antipodal planar tapered slot with an elliptical curvature. Rogers RO4003 with 3.38 dielectric constant and 0.508 mm thickness was used as a substrate. A resistive layer of 50 Ω/□ was sprayed at the designated areas at the lower end of the radiating structure in the top and bottom layers to improve the front-to-back ratio, and thus the detection capabilities of the UWB imaging system.

The design objective is to obtain a directive antenna with a compact size, while maintaining the bandwidth requirement of 3.1 to 10.6 GHz. The following design procedure is proposed and utilized in developing the proposed antipodal antenna.

Step 1. Given the lowest frequency of operation (\(f_l\)), thickness of the substrate (h) and its dielectric constant (\(\varepsilon_r\)), the width (w) and length (l) of the antenna structure, excluding the feeder, can be calculated using the following equation [9]:

\[
   w = l = \frac{c}{f_l\sqrt{\varepsilon_r+1}},
\]

where c is the speed of light in free space.

It is worthwhile to mention that (1) indicates that the antenna’s length and width is chosen to be equal to the effective wavelength calculated at the lowest frequency of operation.

Step 2. The radiating structure of the antenna is formed from the intersection of quarters of two ellipses. The major radii (\(r_1\) and \(r_2\)) and the secondary radii (\(r_{1s}\) and \(r_{2s}\)) of the two ellipses are chosen according to the following equation:

\[
   \begin{align*}
   r_1 &= \frac{w}{2} + \frac{w_m}{2}, \\
   r_2 &= \frac{w}{2} - \frac{w_m}{2}, \\
   r_{1s} &= l, \\
   r_{2s} &= 0.5r_2.
   \end{align*}
\]

According to (2), dimensions of the radiating element are chosen such that the far-end distance between the top and bottom radiators is equal to the effective wavelength at the lowest frequency of operation. Length of each of the radiators at the left and the right end of the antenna’s structure shown in Figure 1 is equal to half of the effective wavelength calculated at the lowest frequency of operation.

Step 3. The width of the microstrip transmission feeder (\(w_m\)) to give the characteristic impedance, \(Z_0\), equal to 50 Ω, can be calculated using the following equations [10]:

\[
   w_m = \frac{120\pi}{\sqrt{\varepsilon_r}} \frac{h}{Z_0}
\]

Step 4. A metallization layer, with a 50 Ω/□ surface resistivity, is added to the top and bottom radiating parts. Shape of the resistive layers is chosen to be a quarter of an ellipse with major and secondary diameters equal to \(r_2\) and \(r_{2s}\), respectively.
Step 5. A transition is added to the structure of the antenna. This is required because the antenna’s radiating element shown in Figure 1 is connected to a parallel strip line, which is a balanced transmission line, whereas the antenna is to be connected to the other devices of the imaging system using a suitable coaxial cable, which is an unbalanced transmission line. The transition from the parallel strip line to the microstrip line is shown in Figure 2, which is adopted from the transitions presented in [11]. The strip line, which is located at the top layer, is connected using a tapered transmission line to the microstrip line, while width of the strip line at the bottom layer is gradually increased to form the ground plane required for the microstrip feeder.

3. RESULTS

The ultrawideband antenna designed according to the above mentioned procedure was manufactured using Rogers RO4003C ($\varepsilon_r = 3.38$, $h = 0.506$ mm) as a substrate. Values of the design parameters $w$, $l$, $r_1$, $r_2$, $r_1$, $r_2$, and $w_m$ (shown in Figure 1) are 50 mm, 50 mm, 26 mm, 24 mm, 50 mm, 12 mm, and 2 mm, respectively. A photo for the developed antenna is shown in Figure 3.

Concerning the resistive layers, a parametric analysis using the software Ansoft HFSSv10 indicated that the best performance concerning the bandwidth and the front-to-back ratio can be achieved when the resistivity of the added resistive layer is in the range from 50 to 100 $\Omega$ $\square$. The lower value was used because of the availability of the 50 $\Omega$ $\square$ chemical mixture to the author.

The validity of the proposed design methodology is verified using the commercial software package, Ansoft HFSSv10, and experimental tests by using a vector network analyzer.

Figure 4 shows the simulated and measured return loss of the manufactured antenna. As can be seen from Figure 5, the 10 dB return loss bandwidth extends from 3.1 GHz to more than 11 GHz covering the required UWB band of 3.1 GHz–10.6 GHz. The simulated result closely resembles the measured result validating the design procedure of the antenna.

The far-field radiation patterns of the antenna were calculated using the software HFSS. They are shown in Figure 5 for the frequencies 4 GHz, 6 GHz, 9 GHz, and 11 GHz. The antenna shows directive properties with an average front-to-back ratio which is greater than 13 dB across the whole band, making it a good candidate for microwave imaging applications. It is worthwhile to mention that without the use of the resistive layers, the front-to-back ratio is around 10 dB.

The measured gain of the antenna is shown in Figure 6, which reveals a moderate gain antenna. The gain is equal to 4.3 dBi at 3 GHz and it increases with frequency till it becomes 10.8 dBi at 10.6 GHz. It is to be noted that the gain measurements were done in comparison with a reference-gain antenna which is the corrugated horn antenna in this case.

As the use of the resistive layer can be responsible for the reduction in the radiation efficiency, suitable calculations with the help of the software HFSS were performed with respect to this parameter. From Figure 6, it is apparent that despite the use of the resistive layers to minimize the backward
radiation (and hence enhance the front-to-back ratio), the proposed antenna has a good efficiency, which is more than 80% across the whole band. This performance is superior in comparison with the antennas reported for use in a microwave imaging system, where 47% efficiency was noted [8].

The time-domain performance of the proposed antenna was also measured. A narrow pulse was synthesized in the network analyzer using the discrete Fourier transform module of the device. The pulse was synthesized after assuming that its frequency spectrum is a rectangular function that extends from 3.1 to 10.6 GHz. Shape of the resulted synthesized pulse is shown in Figure 7. Two copolarized antennas were separated by a distance of 50 cm and the results of the measurement are shown in Figure 7. Note that the excited pulse and the received pulse are normalized with respect to their peak values. The figure reveals that the pulse duration of the antenna is 0.6 nanoseconds. The pulse distortion occurs at the 0.15 level with respect to the peak level of 1, and thus it is almost negligible. The observed results indicate that the developed antenna supports distortionless narrow pulse which makes it an excellent radiator for the purpose of a microwave imaging with high resolution.

As the antenna is to be put on or near the human body, specifically the breast for the case of breast-cancer detection, a study of the effect of the distance from the skin to the antenna on its return loss is investigated. The electromagnetic model used to simulate the breast contains two layers: the first layer is the skin layer with thickness = 2 mm, dielectric constant 36, and conductivity = 4 S/m. The second layer is the breast tissue, which extends to a width of 10 cm, with a dielectric constant = 9 and conductivity = 0.4 S/m [12]. Results of simulation using the software HFSS are shown in Figure 8 for two different distances between the antenna and the human body. Figure 8 indicates clearly that the antenna maintains its ultrawideband performance in spite of being very close to the human body.

The imaging system in which the antenna is to be used contains an array of antennas. Hence, it is important to investigate the value of the mutual coupling between these antennas. The mutual coupling between two identical antennas at different frequencies was calculated using the software HFSS. In the calculations, two antennas were assumed to be parallel to each other and the distance between them was changed. The mutual coupling was calculated at each distance and the results are shown in Figure 9. These results show that the coupling decreases as the distance between the two antennas increases. For a certain distance between the two antennas, the mutual coupling is less for a higher frequency. This is because increasing the frequency means a lower wavelength. Therefore, the distance between the coupled antennas relative to the wavelength is larger. The results depicted in Figure 9 reveal that the mutual coupling between the neighboring antennas at any frequency within the ultrawideband range is less than −20 dB when the distance between the antennas is more than half a wavelength.
It is also important to study the distortion when the radiated pulse propagates through the human body, that is, the skin and the breast tissue in the case of the breast-cancer-detection system. The antenna fidelity is used as an indication of that distortion. The fidelity factor is the maximum magnitude of the cross correlation between the observed pulse at a certain distance and the excitation pulse [13]. The finite difference time-domain method was used for this purpose [14]. In order to reduce the computation domain, Berenger’s perfectly matched layer (PML) is applied as an absorbing boundary condition [15]. To include the frequency dependence of the dielectric constant \( \varepsilon \) and the conductivity \( \sigma \) of the breast tissue over the UWB, the first-order Debye dispersion model was applied [12]:

\[
\varepsilon_i = \frac{j\sigma_0}{2\pi f \varepsilon_0} + \frac{\varepsilon_\Delta - \varepsilon_\infty}{1 + j\frac{2\pi f \tau}{\varepsilon_\infty}}
\]

where \( \tau \) is the relaxation time, and \( \varepsilon_\Delta, \varepsilon_\infty, \text{ and } \sigma_\Delta \) are the Debye model parameters which were selected according to the published data for the breast tissues [12]: normal tissue: \( \varepsilon_\Delta = 10, \varepsilon_\infty = 7, \tau = 7 \text{ ps}, \sigma_\Delta = 0.15 \text{ S/m}, \text{ tumor: } \varepsilon_\Delta = 54, \varepsilon_\infty = 4, \tau = 7 \text{ ps}, \sigma_\Delta = 0.4 \text{ S/m}. \) For the skin: \( \varepsilon = 36, \) and \( \sigma = 4 \text{ S/m}. \)

The result is shown in Figure 10 where the effect of all the scattered/reflected signals is included. It indicates that as the signal propagates through the human body, the fidelity factor decreases. This indicates an increasing pulse distortion inside the human body. For the antenna presented in this paper, the fidelity factor is within reasonable values (more than 70%) even inside the human body.

4. CONCLUSION

The design of a directive ultrawideband antenna for use in a microwave imaging system has been presented. To minimize the backward radiation, the antenna uses resistive layers behind its conductive radiation layers. The simulated and measured characteristics of the antenna have shown that it covers the band from 3.1 GHz to more than 11 GHz. It has a radiation efficiency of more than 80%, which is higher than the recently reported other UWB planar antennas employing resistive layers for microwave imaging applications. The characteristics of the antenna when operating near a human body have been investigated. The simulated results have shown that the antenna maintains its ultrawideband performance concerning the return loss even with the presence of the human body in proximity with the antenna.

The time-domain performance of the antenna has also been studied. It has been shown that the proposed antenna...
has the ability to send and receive very short pulses in a distortionless manner. It has been shown that although the fidelity factor decreases as the signal propagates through the human body, the value of that factor is still within acceptable limits.

The mutual coupling between two identical antennas has been simulated as the antenna elements are used within an array in the microwave imaging systems. It has been shown that the mutual coupling is less than −20 dB when the distance between the neighboring antennas is more than half a wavelength.

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Miniaturized Microstrip-Fed Tapered-Slot Antenna With Ultrawideband Performance

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Abstract—A method to design a microstrip-fed antipodal tapered-slot antenna, which has ultrawideband (UWB) performance and miniaturized dimensions, is presented. The proposed method modifies the antenna’s structure to establish a direct connection between the microstrip feeder and the radiator. That modification, which removes the need to use any transitions and/or baluns in the feeding structure, is the first step in the proposed miniaturization. In the second step of miniaturization, the radiator and ground plane are corrugated to enable further reduction in the antenna’s size without jeopardizing its performance. The simulated and measured results confirm the benefits of the adopted method in reducing the surface area of the antenna, while maintaining the ultrawideband performance.

Index Terms—Miniaturization, planar antenna, tapered-slot antenna, ultrawideband (UWB) antenna.

I. INTRODUCTION

TAPERED-SLOT ANTENNAS (TSAs) offer a wide operational bandwidth and high gain. Therefore, they are widely used in radar, remote sensing, and ultrawideband (UWB) communications.

TSAs are designed using different types of tapering, such as linear, constant width, exponential (Vivaldi), broken linear, dual exponential, and elliptical [1]. In its basic configuration, the TSA is fed by a slotline with high input impedance (≈ 50Ω). To overcome this problem and achieve the required matching with the widely used 50Ω microstrip feeder, several types of feeding arrangements that use transitions and/or baluns were developed [1]. However, those feeding structures resulted in a larger size and a lower efficiency due to the additional insertion losses introduced by the utilized feeding configurations.

To minimize the feeding problems of the TSAs, the microstrip-fed antipodal structure was proposed [2]. In the antipodal arrangement, one of the radiating fins is converted to form a transition and a microstrip line, while the other one is shaped to create a tapered ground plane for the microstrip. Design guidelines have recently been proposed to build microstrip-fed antipodal TSAs [3]. However, the utilized feeding structure, which includes slotline-to-microstrip transition and tapered ground, introduces an additional size to the antenna.

The large dimensions of the TSA represent a serious challenge toward its use in the new generations of highly compact wireless communication systems, which offer a limited space for the RF front-end including the antenna. The relatively large size of the TSA originates from two factors: first, the use of a complicated feeding structure, and second, the design requirement that the width of the taper’s opening and its depth should not be less than one wavelength calculated at the lowest frequency of operation [4], [5].

In this letter, the antipodal structure of the TSA is miniaturized following two steps. First, the antenna’s configuration is modified to enable direct connection with a microstrip feeder. Second, corrugated structures are utilized in the radiator and ground plane.

II. DESIGN

Configuration of the conventional antipodal TSA is shown in Fig. 1(a). Using the design method of [3], it is possible to find that the most compact overall dimensions for the antenna to achieve an UWB bandwidth (3.1–10.6 GHz) are \( l = 41\text{mm} \) and \( l' = 21\text{mm} \). It is assumed that an elliptical tapering is used and Rogers RT6010 (thickness = 0.812 mm, dielectric constant = 10.2) is the substrate.

In the first step of the miniaturization process, the structure is modified by removing the tapered ground plane and the slotline-to-microstrip transition. A microstrip line is connected directly to the top layer in the manner shown in Fig. 1(b), whereas the bottom layer acts as a ground plane. The slot \( (s) \) between the top and bottom layer is used to achieve a perfect matching between the microstrip feeder and the radiator. With this modification in the structure, the overall dimensions of the structure become \( l = 41\text{mm} \) and \( l' = 21\text{mm} \), which means a reduction in size by 33%. The optimized values for the other design parameters \( (s \) and \( l'c) \) are 0.3 and 40 mm, respectively.

It is worthwhile to mention that the microstrip feeder in Fig. 1(b) is curved away from edge of the structure to ease the connection of the feeder to the external port using a Subminiature A (SMA) connector and to prevent the unwanted radiation from the microstrip line in case it extends along edge of the structure.

In the second step of the miniaturization process, symmetrical corrugations are made in the radiator and the ground plane in the manner shown in Fig. 1(c). This step helps the designer reduce the length \( (l) \) and width \( (l') \) of the antenna while maintaining the ultrawideband performance. The depth of the corrugations is chosen to be less than a quarter of the effective wavelength at the lowest frequency of operation (3.1 GHz) so that the corrugations present an inductive reactance to the passing wave [6].


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Fig. 1 Configuration of (a) the traditional antipodal TSA, (b) modified structure, and (c) miniaturized structure.

This added inductance increases the electric length of the structure. Therefore, the corrugated antenna’s structure resonates at a lower frequency in comparison with a noncorrugated structure having the same dimensions [7]. The initial values for the depth and width of the corrugations are found using the principles mentioned in [7], whereas the final values are found using the optimization capability of CST Microwave Studio.

III. RESULTS AND DISCUSSION

To validate the presented method, the miniaturized antenna’s configuration [Fig. 1(c)] was manufactured. A photograph of the manufactured antenna is shown in Fig. 2. Values of the design parameters \( l, r, s, f, s, \) and \( c \) obtained after optimization using Microwave Studio are 25, 30, 0.5, 0.3, and 18 mm, respectively. Depth of the corrugations ranges from 1 mm for the inner corrugation to 3 mm for the outer corrugation, while they have uniform width and separation that are equal to 0.5 mm.

It is clear from values of the final design parameters that the overall dimension of the antenna (25 mm \( \times \) 30 mm) represents only about 20% of the antenna’s surface area designed using the traditional approach for UWB antipodal TSA (60 mm \( \times \) 60 mm).

The developed antenna was tested via simulations and measurements. During measurements, the cable connecting the antenna to the measuring instrument was embedded in an absorbing sheet to avoid any interaction between the near-field of the antenna and the cable. The simulated and measured return losses of the miniaturized antenna together with the simulated results for the original full-size structure of Fig. 1(a) and the modified noncorrugated structure of Fig. 1(b) are shown in Fig. 3. The full-size, modified, and miniaturized antennas have UWB performance with bandwidths extending from 1.3 to more than 11 GHz, 1.5 to more than 11 GHz, and 1.8 to 10.8 GHz, respectively, assuming the 10-dB return loss as a reference. This result indicates that despite the huge size reduction, the adopted miniaturization method maintained the UWB performance of the antenna. A good agreement can be seen in Fig. 3 between the simulated and measured results.

Concerning the radiation pattern of the antenna, the measured results at the two main planes \( (x, y) \) and \( (y, z) \) and different frequencies depicted in Fig. 4 indicate a directive performance. The antenna has end-fire properties as the main beam is in the axial direction of the tapered slot (\( r \)-axis), i.e., at \( \varphi = 0 \) as shown in the \( r, \varphi \)-plane and \( \varphi = 0 \) as shown in the \( r, \varphi \)-plane. The cross-polarized field was also measured and found to be less than the copolarized field by more than 10 dB across the whole band of operation and angles of radiation. The measured gain of the antenna confirms its directive properties. The results shown in Fig. 5 reveal a gain that is between 2.7 and 8 dBi across the
ultrawideband. It is to be noted that the huge size reduction of
the antenna due to miniaturization has only resulted in a modest
reduction of 1 dB in the gain compared to the measured gain for
the traditional antipodal TSA [3].

The other important parameter for the UWB antennas, es-
pecially when used to send/receive pulsed signals, is the time
domain response. For this purpose, two miniaturized antennas
were put at a distance of 30 cm such that their tapered slots face
each other. A UWB pulse synthesized in the vector network an-
alyzer to cover the band 3.1–10.6 GHz was transmitted from
the first antenna, and the received pulse by the second antenna
was measured. The result is shown in Fig. 6, where amplitudes
of the transmitted and received pulses were normalized to have
a peak equal to 1. It is clear that the antenna is efficient since
the received pulse has low levels of distortion and ringing. The
low-distortion time domain performance of the miniaturized an-
tenna was also confirmed by calculating the fidelity factor of
the antenna using the method presented in [8]. It was found to
be larger than 95% for the face-to-face operation, which is the
normal case for directive antennas.

IV. CONCLUSION

A method has been presented to design miniaturized an-
tipodal tapered-slot antennas with ultrawideband performance.
The proposed miniaturization process includes two steps. First,
the antenna’s structure is modified to enable a direct connection
to a microstrip feeder, and second, corrugated structures are
used in the top and bottom layers of the antenna. The simulated
and measured results in the frequency and time domain have
confirmed the success of the proposed design approach.

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better performance at the rest of the UWB. Concerning the return loss and isolation, the elliptical shape has a better performance across most of the UWB.

4. CONCLUSION

In this article, a theoretical model has been used to study effect of the tapering shape on performance of the broadband high directivity microstrip reflectarray across the ultra wideband range. Simulations and measurements have been used to verify the findings of the model.

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It is clear from this equation that, for certain conductivity, increasing the thickness (t) reduces value of the resistance, and hence decreases the conductor loss. According to Figure 2, the total losses is around 0.8 dB when t = 0.035 mm, while it is <0.3 dB when t = 0.5 mm.

To confirm effect of t on the conductor loss and to show the relative values of the dielectric and conductor loss separately, the previous simulation was repeated, but in this case the dielectric loss is included when dielectric tangent loss of the main conductor is zero. The result, which is shown in Figure 2, reveals a similar trend to the case when the dielectric loss is excluded, but with lower values for the losses as only the conductor loss is included. Comparing the two cases of Figure 2 shows that the conductor loss is the main contributor in the total losses of the microstrip reflectarray when thickness of the conductive coating is very small. The dielectric loss can be estimated to be around 0.2 dB, in general, for all the studied situations, whereas the conductor loss for the case t = 0.035 mm is around 0.6 dB. This proves the importance of tackling the conductor loss by choosing the suitable thickness for the conductive coating. This would eventually result in a high efficiency for the microstrip reflectarray.

Concerning the phase performance of the unit cell under investigation, it is clear from Figure 3 that increasing thickness of the conductive coating causes the phase slope variation to be slower, which means a broader bandwidth, while it has a limited effect on reducing the phase range. The simulation was repeated assuming this time that the dielectric loss tangent is zero, and similar results were obtained. It is to be noted that for the case t = 0.5 mm, the phase continues to change after the 11 GHz, which is the end mark of Figure 3, till it becomes close to value of the phase at the other cases. It is also worthwhile to mention here that if it is required to cover a wider phase range, then it is possible to use a double cross shaped rings and a thick substrate of foam with dielectric constant close to 1 [6].

To verify the results obtained via simulations, the unit cell shown in Figure 1 was manufactured and tested using a vector network analyzer. Two samples, which have a different thickness of conductive coating, were developed: The first sample has t = 0.035 mm, whereas the second sample has t = 0.07 mm. During testing, each of the developed samples was inserted inside an
X-band waveguide, which is connected to the network analyzer using a standard waveguide-to-coaxial cable termination. The measured amplitude of the return loss for the two samples is shown in Figure 4. The measured results depicted in this figure confirm the previous conclusions from simulations; increasing thickness of the conductive coating decreases the losses and shifts the resonant frequency to a higher value. The total losses at resonance is equal to 1 dB and the resonant frequency is equal to 9.8 GHz for the case $t = 0.035\ \text{mm}$, whereas the total losses and the resonant frequency are equal to 0.7 dB and 9.9 GHz, respectively, for the case $t = 0.07\ \text{mm}$. Variation of the measured phase of the developed samples with frequency is depicted in Figure 5. It is confirmed from this figure that increasing thickness of the conductive coating shifts the resonant frequency to a higher value with almost the same phase range.

Comparing the measured results, which are shown in Figures 4 and 5, with the simulated results, which are shown in Figures 2 and 3, reveals that their general variation is very similar. However, the measured losses are higher by about 0.2 dB, whereas the resonant frequencies are lower by about 0.3 GHz, compared with their simulated counterparts. These small differences between the measured and simulated results can be referred to the possible leakage due to the difficulty in fixing the developed samples at the bottom of the waveguide, besides the additional losses associated with the used termination and connectors.

3. CONCLUSION

In this article, effect of the conductive coating thickness on performance of the microstrip reflectarray has been investigated. A cross shaped ring designed to operate at the X-band was used as a unit cell. The simulated and measured results have shown that increasing thickness of the conductive layer decreases the losses significantly, reduces the phase slope and shifts the resonant frequency to a higher value, while it has a negligible effect on the phase range. The results of this article have revealed the importance of choosing a suitable conductive layer thickness for a high efficiency of the microstrip reflectarray.

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three-way DPA using two-stage GaN HEMT cells can be a promising solution for repeater systems.

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COMPACT DIRECTIONAL ANTENNA FOR ULTRA WIDEBAND MICROWAVE IMAGING SYSTEM

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ABSTRACT: A planar antenna of tapered slot configuration for use in ultra wideband microwave imaging systems aimed for early breast cancer detection is presented. It is designed to operate across the ultra wideband frequency (3.1–10.6 GHz) in a liquid of a high dielectric constant that matches the electric properties of average breast tissues. It has a very compact size with overall dimensions of 0.9 cm × 1 cm. The antenna’s ultra wideband performance even when it is close to the breast tissues and its distortionless response in the time domain make it suitable for the microwave imaging systems utilizing a short-pulse radar technique. © 2009 Wiley Periodicals, Inc. Microwave Opt Technol Lett 51: 2898–2901, 2009; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24764

Key words: ultra wideband antenna; planar antenna; microwave imaging; breast cancer

1. INTRODUCTION
Ultra wideband (UWB) microwave imaging is a promising method for some medical applications such as breast cancer detection. The reason is that it offers good penetration and high resolution characteristic of the acquired image. The possibility of using this active microwave technique to detect breast cancer stems from the significant contrast in dielectric properties between normal and cancerous tissue [1].

In this UWB imaging system, a very narrow pulse is transmitted from an antenna to penetrate the breast. The scattered signal due to different layers of the breast tissues is collected by array of antennas surrounding the breast and then a suitable signal processing algorithm is applied to investigate the existence of any cancerous tissues. To achieve a high resolution over a large dynamic range, the transmitting/receiving UWB antenna should be compact in size and directive, with distortionless pulse performance.

Many types of UWB antennas were proposed to be part of the microwave imaging systems [2–5]. The tapered slot antennas [2–4] seem to be the best candidate for imaging systems in terms of bandwidth, gain, and impulse response. Also, they are convenient to form arrays. However, many designs available in the antenna’s literature lead to a large size to achieve the required ultra wideband performance.

In the presented work, a UWB planar antenna featuring a very compact size with dimensions of (0.9 cm × 1 cm) is described. The simulated performance of the proposed antenna in an environment, which is similar to the real situation in the breast imaging case, shows an ultra wideband behavior with a distortionless pulse operation. The UWB performance of the antenna is confirmed via measurements.

2. ANTENNA DESIGN
A microwave imaging system for breast cancer detection is assumed to include a circular array of antennas that encircle the breast. This array is immersed in a liquid with a high dielectric constant to achieve the best possible matching with the breast tissues [4]. This is required to reduce the reflected/scattered signals at the skin layer interface, and thus increase the dynamic range of the imaging system.

Configuration of the antenna that is proposed for use in the array is shown in Figure 1. It uses Rogers RT6010 (εr = 10.2, thickness = 0.64 mm) as a substrate. The antenna is assumed to be immersed in a liquid with a high dielectric constant to achieve a good matching with the breast tissues [6]. The overall dimension of the antenna (the length L and the width W in Fig. 1) is chosen initially to be equal to half of the effective wavelength calculated at the centre frequency of operation (6.85 GHz). The radiator, which is located at the top layer of the substrate, and the ground plane at the bottom layer are in the form of quarter an ellipse with a major diameter equal to W. There is a slot S between the radiator and the ground plane. The initial value for this slot is chosen to be equal to the substrate’s thickness. The secondary diameter of the quarter ellipses representing
the radiator and the ground plane is initially chosen to be equal to the major radius minus the slot \( S \). With this choice for the antenna dimensions, the overall size is very compact; however, it will have an inferior performance at the low frequency band (around 3 GHz). To improve the performance at that band, pairs of symmetrical slots are cut from the radiator and the ground plane in the manner shown in Figure 1. Those slots increase the path length of the current near the end of the structure. This makes the path length of the current effectively larger than the physical length. This eventually improves the performance at the lower end of the band without a significant effect on the rest of the covered band.

The initial antenna’s dimensions are optimized using the software CST Microwave Studio. The final dimensions are: \( L = 10 \) mm, \( W = 9 \) mm, \( W_s = 0.4 \) mm, \( L_s = 2 \) mm, \( d_s = 3 \) mm, \( S = 0.5 \) mm. The antenna is fed using a microstrip line with width equal to \( W_f = 0.45 \) mm.

3. RESULTS AND DISCUSSIONS

Performance of the proposed antenna was first verified via computer simulations. Next, a prototype was manufactured and tested to confirm its simulated performance.

In the imaging system for breast cancer detection, the antenna is to be at a close distance from the breast. Therefore, effect of the breast tissues on the antenna’s performance is investigated. The electromagnetic model used to simulate the breast contains two layers: The first layer is the skin layer with thickness = 2 mm, dielectric constant = 36, and conductivity = 4 S/m. The second layer is the breast tissue, which extends to a width of 10 cm, with a dielectric constant = 9 and conductivity = 0.4 S/m [7]. Results of the simulation are shown in Figure 2 for different distances between the antenna and the breast. Figure 2 indicates clearly that the antenna maintains its ultra wideband performance despite being very close to the breast tissues.

The far-field radiation pattern of the antenna was calculated and it is depicted in Figure 3 at 3 and 10 GHz. The antenna...
shows directive properties with an average front-to-back ratio which is greater than 10 dB making it a good candidate for microwave imaging applications. The imaging system in which the antenna is to be used contains an array of antennas surrounding the breast. Thus, it is important to investigate value of the mutual coupling between those antennas. The mutual coupling between two identical antennas at different frequencies was calculated assuming two values for the distance between them. The two mutually coupled antennas were assumed to be parallel to each other in the manner shown in Figure 4. Results of the calculation are shown in Figure 4. It shows that the coupling is less than 25 dB across the whole ultra wideband when the distance between the two antennas is 20 mm, which is around half of the effective wavelength at the lowest frequency of operation which is 3.1 GHz. Figure 4 also shows that the coupling decreases as the distance between the two antennas increases.

The time domain performance of the proposed antenna was also calculated. A narrow pulse was assumed to be transmitted from one antenna and the received pulse was calculated at a copolarized receiving antenna which is at a distance of 30 cm from the transmitter. The pulse shape was chosen such that it contains the UWB frequency spectrum of 3.1 to 10.6 GHz. Shapes of the transmitted and received pulses are shown in Figure 5. Note that the excited pulse and the received pulse are normalized with respect to their peak values. The figure reveals that the pulse distortion occurs below the 0.2 level with respect to the peak level of 1, and thus it is almost negligible. The observed result indicates that the designed antenna supports distortionless narrow pulse operation which makes it an excellent choice for the purpose of a microwave radar imaging.

To confirm the UWB performance of the antenna, a prototype was manufactured and tested. Figure 6 shows the measured return loss of the antenna when immersed in a liquid, which is a mixture of distilled water and saline with a dielectric constant of around 10 at 3 GHz. As can be seen from this figure, the 10 dB return loss bandwidth extends from 2.1 GHz to more than 11 GHz covering the required UWB band (3.1–10.6 GHz). There is a good agreement between the measured and simulated results shown in Figure 6 till a frequency of 10 GHz. After that, there is a significant difference between them, and this difference could be due to variation of the dielectric constant of the liquid used in the measurements with frequency.

4. CONCLUSION

The design of a planar tapered slot ultra wideband antenna for use in a UWB microwave breast cancer imaging system has been presented. The antenna has a compact size of 0.9 cm × 1 cm. To improve the matching between the antenna and the breast tissues, the antenna is immersed in a liquid of a high dielectric constant. The simulated results have shown that the antenna covers the ultra wideband range even when it is very close to the breast with a low distortion in the time domain performance and a low mutual coupling between closely spaced antennas. The measured characteristics of the antenna have confirmed its ultra wideband operation.
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MINIMUM LOSS CONDITION OF A BENT RECTANGULAR HOLLOW WAVEGUIDE
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ABSTRACT: The leakage loss of a hollow dielectric waveguide is analyzed analytically and numerically. A minimum loss condition of a bent rectangular hollow waveguide is derived in terms of the refractive index of the cladding using the perturbation method. The validity of the derived minimum loss condition is confirmed by the beam-propagation method. © 2009 Wiley Periodicals, Inc. Microwave Opt Technol Lett 51: 2901–2902, 2009; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24761

Key words: beam-propagation method; perturbation method; leakage loss; bend loss

1. INTRODUCTION
A hollow dielectric waveguide has attracted much attention owing to its unique propagation characteristics [1–6]. The loss of a straight and a bent slab hollow waveguide have already been investigated in detail using the perturbation method [2, 3]. It has also been indicated that the loss of a straight rectangular hollow waveguide is approximated by the sum of the two losses for the TE and TM modes in a two-dimensional (2D) slab hollow waveguide [3, 4]. These facts motivate us to apply the loss formulae derived by the perturbation method for the slab hollow waveguide to a bent rectangular hollow waveguide.

In this article, we explicitly show the loss formula of a bent rectangular hollow waveguide by the perturbation method. Taking advantage of the analytical technique, we will derive a minimum loss condition in terms of the refractive index of the cladding. We also analyze a straight and a bent rectangular hollow waveguide by the three-dimensional imaginary distance beam-propagation method (3D-BPM), which demonstrates the validity of the derived minimum loss condition.

2. FORMULATIONS
We study a rectangular hollow waveguide composed of a low-refractive index material surrounded by a high-refractive index material, as shown in Figure 1(a). The refractive indices of the core and cladding are designated as n_{co} and n_{cl}, and the core width and height are as 2w and 2h, respectively. The hollow waveguide is bent with a radius R in the x–z plane. It has been found that the leakage loss in a straight rectangular hollow waveguide can be estimated by the sum of two losses obtained from straight slab hollow waveguides in the TE and TM modes [3, 4]. It is, therefore, expected that the loss in a bent rectangular hollow waveguide can be approximated by the sum of two losses obtained from the perturbation method [2, 3] for the straight and the bent slab configurations shown in Figures 1(b) and 1(c), respectively. The sum of the two losses leads to the following attenuation constant for the E'_{mn} mode:

\[ \gamma = \frac{A + BC}{(n_{co} \delta n)} \sqrt{C} - 1 \]  

where k_{0} is the free-space wavenumber, A = w_{0}^{2}/w^{3}, B = w_{0}^{2}/w^{2}, C = (n_{co}/n_{cl})^{2}, u_{m} = m\pi/2, and u_{n} = n\pi/2, in which m and n are mode numbers. For the E'_{mn} mode, we can obtain \gamma_{mn}' by interchanging A and B in Eq. (1). In this article, the order of the fundamental mode is defined as one. The coefficient \gamma includes the effect of bending loss, and becomes unity for a straight waveguide. Depending on the bending radius R, \gamma is evaluated by [2, 3],

\[ \gamma_{L} = 1 - \frac{2}{\pi} \left( 1 - \frac{15}{4\mu_{m}} \right) D^{2}E^{2} - \frac{5}{9} \left( 1 - \frac{105}{2\mu_{m}} + \frac{495}{4\mu_{m}}^{2} \right) D^{4}E^{4} \]  

for large R (2)

\[ \gamma_{S} = DE \]  

for small R (3)

where D = (n_{co}k_{0}w/u_{m})^{3} and E = w/R. In the derivation of Eq. (1), it is assumed that the core width and height are sufficiently large compared with the wavelength.

It is interesting to note that for a straight (\gamma = 1) rectangular hollow waveguide, Eq. (1) is substantially the same as that derived by Isaac and Khalil using a ray-optics approach [6]. In other words, Eq. (1) can be regarded as an equation extended to a bent rectangular hollow waveguide.

To find the location of the local extremum of the loss, we differentiate Eq. (1) with respect to the refractive index of the cladding, that is, \partial\gamma_{mn}'/\partial n_{cl} = 0. As a result, we obtain the following minimum loss condition:

\[ n_{cl}^{\text{min}} = n_{co}/\sqrt{2 + F} \]  

where F = (n/\mu)^{2}(w/h)^{2} \gamma_{L} for the E' mode and F = F^{-1} for the E'' mode.

For a straight square (h = w) waveguide with m = n, the minimum loss condition of both E' and E'' modes is simply expressed as

\[ n_{cl}^{\text{min}} = \sqrt{3}n_{co} \]  

which serves as a rough guideline for determination of n_{cl}, as can be seen in Figures 2 and 3.
Strain Imaging of the Breast by Compression Microwave Imaging

A. Abbosh and S. Crozier

Abstract—A method that uses microwave pulses to achieve strain imaging of the breast is presented. In the proposed method, the breast is inserted in an enclosure that defines the boundary conditions for the breast deformation under the influence of external pressure. The upper plate of the enclosure, which also includes an antenna array, is attached to a compression tool, whereas the lower and the front plates are fixed. The breast is allowed to extend in the lateral direction when pressed by the top plate. Each of the antennas at the top plate is used to send an ultrawideband pulse to penetrate the breast and measure the backscattered pulse. Two sets of measurements are taken: one pre- and another post-compression of the breast. A sliding window of cross correlation is then performed on the two sets of data to establish the time delay of each segment of the scattered pulse due to compression. That time delay is then employed to get a three-dimensional strain image of the breast. As lesion tissue is typically much stiffer than normal breast tissue, then regions of low or zero strain indicate areas in need of further diagnostic checks. Full-wave simulations are used to validate the presented imaging method.

Index Terms—Breast cancer, microwave imaging, strain imaging.

I. INTRODUCTION

DIFFERENT modalities for breast cancer detection are being investigated around the world. One of those modalities is microwave imaging [1]–[6], which relies on the assumption of a high dielectric contrast between healthy and malignant breast tissues. However, it has been shown recently [7] that this is not always the case. For example, the fibro glandular tissues have a dielectric constant that is very close to that of malignant tissue, such that the contrast between them is as low as 1.2. In many cases, the lesions are embedded within fibro glandular tissue. With a very low surrounding contrast, it would be extremely hard to achieve a successful detection by relying on the electrical properties alone. Therefore, another dimension to the microwave imaging modality is needed.

The changes in the mechanical and electrical properties of tissues are directly related to their pathological state [8]. Tumors in the breast are stiffer with higher elasticity modulus compared to the healthy tissues [9]. Thus, manual palpation is the first diagnostic method used to investigate breast pathology. Unfortunately, manual palpation is unsuccessful in many cases due to the small size of the lesion or/and its deep location.

There has been extensive research on the possibility of utilizing the elasticity contrast between tissue types using ultrasound techniques [9]. However, the success is limited as some forms of breast lesions do not have the echogenic properties that make them detectable by ultrasound.

The approach presented in this letter can be defined as an automated palpation via using ultrawideband pulses to measure the contrast in the stiffness of the healthy and malignant tissues. To achieve that target, a controlled compression is applied to the breast, then pre- and post-compression backscattered ultrawideband (UWB) signals are cross-correlated to obtain a three-dimensional strain image of the breast.

II. PROPOSED METHOD

The principle of the presented method is that tissue compression (stress) produces displacement (strain) within the tissues and that the displacement is smaller in harder tissue than in softer tissue. Therefore, by measuring the tissue strain induced by compression, it is possible to estimate tissue hardness, which may be useful in differentiating normal and abnormal tissue types, including malignancies. To this end, the breast is inserted between two horizontal plates as shown in Fig. 1. The top plate is connected to a controlled compression tool, and it carries a two-dimensional array of antennas. To protect the antennas from effect of compression and to achieve a perfect matching between the antennas and the breast tissues, a superstrate layer that has a dielectric constant equal to the average dielectric constant of the breast tissues is used. The lower plate is fixed to support the breast. There is a vertical plate in front of the breast to limit the breast extension to only the lateral direction, i.e., z-axis in this model. The bottom and front plates are assumed to have a dielectric constant equal to that of the skin and covered with absorbing material to minimize any unwanted back-reflection from the boundaries.

A UWB pulse is transmitted from any of the antennas at the top plate, and the backscattered signal is measured at the same antenna. This step is repeated for all the antennas that form the

Fig. 1. Schematic diagram showing how the proposed system is used.
two-dimensional array in order to scan the whole breast. The top plate is then pressed against the breast in a carefully calculated manner. The measurement is repeated. The signals collected in the first set are cross-correlated with the second set of measurements. The cross correlation is calculated along the whole axial line between the top and bottom plates by utilizing a sliding time window that covers the whole backscattered signal.

If the axial line includes only healthy tissues, the cross correlation will reveal a uniform displacement of tissues along that line by a certain amount that depends on the elasticity modulus of the healthy tissues and the amount of pressure applied. If there is any tumor at that line, part of the line that includes the tumor will show a very low displacement variation, or zero strain. By repeating this operation at all the possible axial lines, a three-dimensional image that shows the strain distribution inside the breast can be obtained.

The axial distribution of strain throughout the breast volume is estimated according to the following summarized steps.
1) Get a set of measurements for the backscattered signal before compression. The backscattered signal before compression is divided into time windows; the size of those windows depends on the required resolution.
2) Compress the breast in the longitudinal direction by a certain distance $\lambda_i$ that depends on the required strain using a controlled compression plate, and acquire another set of measurements for the backscattered signal. The backscattered signal after compression is divided into time windows using the same size of those used for the signals before compression.
3) Cross correlation is performed to find the time shifts between each pair of windows from the two sets of measurements. The maximum cross correlation achieved between any pair of segments from the signals before and after compression means a high similarity between the two signals, an indication of a scattering by the same part of breast. The amount of time shift between the two segments is calculated depending on the position of the relevant windows.
4) The local strain $\varepsilon(i)$ at a certain depth that represents the segment of data $i$ covered by the window is calculated

$$\varepsilon(i) = c (\mathcal{H}(i) - \mathcal{H}(i-1))/\mathcal{H} \sqrt{\varepsilon_r},$$

where $\mathcal{H}(i)$ is the time shift between $i$th segments in the data pair, $c$ is speed of light, and $\varepsilon_r$ is the average dielectric constant of breast tissues.
5) The time window is shifted to cover another segment of data, and the calculation is repeated to find the local strain of another object within the breast.

This process is repeated for all the segments in the backscattered signals. The result is the distribution of strain in the breast. The part that shows a sudden change to low or zero strain indicates a suspicious tissue element.

III. MODELING

The biomechanical breast models give distribution of Young’s ($E$) and Poisson’s ($\nu$) modulus for the main tissues of breast (fatty, glandular, cancerous, and skin tissue) across a certain range of strains. Young’s modulus is defined for a certain material that is under the effect of a normal stress ($\sigma$), which results in a normal strain ($\varepsilon$) as [10]

$$E = \varepsilon / \delta \varepsilon,$$

where $E$ is the Young’s modulus. For low strain values in the proposed imaging technique for a high contrast in the mechanical properties. Moreover, low values for the strain means that the assumption of isotropic breast model is valid as explained earlier. However, using very low values for strain means an almost impractically low deformation. Thus, a compromised value for the strain level around 10% is to be used.

In this letter, the following moderate values of the biomechanical parameters based on measurements using ex vivo breast samples [12] are used: Young’s modulus for fat and glands are equal for low strain values (less than 10%), and they are equal to 3.25 kPa. For high-grade invasive ductal carcinoma tissues, $E = 42.5$ kPa. Hence, the contrast in the utilized biomechanical model is 13. According to [11], the skin can be modeled as a linear tissue with a Young’s modulus of 10 kPa and a thickness of 1 mm. As the skin has a very small thickness and as the malignant tissues are not usually close to the skin, the relatively low contrast (4.25:1) between the malignant tissues and skin has a very limited effect on the strain imaging technique.

In order to test the proposed method, a three-dimensional heterogeneous model of breast is needed to find the distribution of deformation under the effect of a certain external stress. For the system presented in this letter, the breast inside the apparatus can be approximated by the heterogeneous model presented in Fig. 2. The top layer is assumed to be movable to exert a uniformly controlled stress of $\sigma_y$.

The utilized heterogeneous model is formed from 40 layers of fat each with 2 mm thickness. Those layers have electrical and mechanical properties that vary randomly by up to ±50% from the nominal values. The model includes 100 spherical-
shaped glands that are distributed randomly within the model as depicted in Fig. 2. The radius of those glands varies randomly between 1–8 mm. The electrical and mechanical properties of those spherical glands vary by up to ±50% from the nominal values. The overall dimension of the utilized model is 100 × 80 × 100 mm³.

The multidisciplinary software ANSYS [13] is used to calculate the deformation due to the applied stress. The applied stress at the top layer is \( \sigma_{\text{E}} = 1.327 \text{kPa} \). This value is carefully chosen in order to make sure that the strain \( (\varepsilon = \frac{\sigma_{\text{E}}}{E}) \) is around 10%. The results of deformation calculated using ANSYS for two cases, due to space limitation, are shown in Fig. 3. After inspecting those figures closely, it is possible to see that the deformation is uniformly distributed across the model except at the tumor, where the deformation variation is perturbed by the tumor.

The three-dimensional models before and after applying the stress are imported in the software CST Microwave Studio to calculate the scattered signals.

IV. STRAIN IMAGING

In order to prove the possibility of using a UWB signal to get useful strain image of breast, a three-dimensional model before applying the pressure (Fig. 2) and after that as generated from Figs. 3 are used in the full-wave electromagnetic software CST Microwave Studio in order to calculate the back-reflected signals. The width of the model in the third dimension (z) is assumed to be 10 cm. In order to include the dispersive characteristics of breast tissues in the utilized model, Debye model is used to find variation of the values of the dielectric constant and conductivity with frequency. According to that model, the electrical properties of breast tissue can be calculated using the following equation [14]:

\[
\varepsilon_r = \varepsilon_\infty + \Delta \varepsilon / (1 + j \tau) + \sigma_n / (j \omega \varepsilon_0)
\]

where \( \varepsilon_\infty, \Delta \varepsilon, \tau, \) and \( \sigma_n \) are the tissue-dependent Debye parameters. \( \omega \) is the radian frequency, and \( \varepsilon_0 \) is the permittivity of free space. The Debye parameters for the different tissues of the breast are assumed as follows [14]:

- Gland: \( \varepsilon_\infty = 13.8, \Delta \varepsilon = 21.5, \tau = 13 \text{ ps}, \) conductivity = 1.74 S/m;
- Fat: \( \varepsilon_\infty = 3.11, \Delta \varepsilon = 1.5, \tau = 13 \text{ ps}, \) conductivity = 1.41 S/m;
- Tumor: dielectric constant = 3.1, conductivity = 4 S/m.

For the skin [15]: dielectric constant = 3.1, and conductivity = 1 S/m.

A corrugated tapered slot antenna [16] was utilized in the simulations. For a perfect matching with the breast tissues, the antenna is assumed to be immersed in the dielectric medium of the top plate of the apparatus shown in Fig. 1. Assume that the dielectric constant of the top layer that includes the antennas is 10 and that Rogers RT6010 (dielectric constant = 10.2, thickness = 0.635 mm) is used as the substrate for the antenna. Using the design rules presented in [16], it is possible to show that the antenna is directive and compact with dimensions of 1.2 × 0.8 cm².

A time-domain pulse of bandwidth 2–12 GHz is transmitted separately from each of the antennas at the top plate. The backscattered signal is collected by the same transmitting antenna before and after compression. The time delay of each segment of the time-domain UWB pulses for pre- and post-compression is calculated, and the strain distribution is estimated using the proposed algorithm for different sections of the model.

The results of the calculations using the proposed method are presented in Fig. 4 for different cases. For the heterogeneous model without a tumor, the result shown in Fig. 4(a) does not include any suspicious area. The strain has a smooth variation. It has a maximum value at the top and decreases gradually till it becomes zero at the bottom. The parts of the model close to the boundaries have zero strain due to the fixed plates and the chest wall.

The results depicted in Fig. 4(b)–(e) present breasts with tumors of different locations and sizes. In those images, the strain has a smooth variation except at the tumor region, where there is a sharp reduction in the strain and it becomes almost zero. This result clearly indicates a very stiff object (suspicious tumor) within the breast at the positions shown.

It is to be noted that, although it might be difficult to detect the presence of a 1-mm tumor in Fig. 4(d) visually, the quantitative strain values calculated using the proposed method reveal that the strain is around 0.5% at the location of the tumor, whereas it is more than 3% in the surrounding region. The high 6:1 contrast in the strain value is an obvious indication for the presence of a suspicious tumor at that location.

For the situation when the tumor is close to the fixed boundaries of the breast, such as close to the chest or the front and bottom plates, the result depicted in Fig. 4(e) is presented. In this case, a 4-mm tumor is located close to the front fixed plate. The fatty and glands areas close to that plate cannot be deformed by a large value, although their Young's modulus is low as they are not free to move. Thus, those healthy tissues display a very low strain in the calculated results, which makes it difficult to visually detect small tumors if they exist within or in close proximity to those healthy tissues. However, it is still possible to recognize the presence of the tumor when using the quantitative data, which reveals a variation of more than 50% of strain in the tumor area compared to that of the surrounding healthy tissues.

It is clear from the presented results that the tumor can be easily detected in all the investigated cases. The position of all the detected tumors is almost the same as that of the assumed tumors, whereas the size appears to be slightly larger mainly due to the elongation in the horizontal direction.
it decreases as the tumor’s position becomes close to the fixed plates and chest wall. This behavior can be clarified by the fact that although the Young’s modulus for the healthy tissues at the lower part of the breast or the front plate and chest wall is still relatively low, the effective stress, and thus the resulting strain, is low due to effect of the boundary conditions, i.e., fixed side and bottom plates.

V. CONCLUSION

A method for the possibility of using ultrawideband signals to get a three-dimensional strain image of breasts has been presented. It has been shown theoretically and via full-wave simulations that the sliding window cross correlation between the backscattered signals before and after breast compression can detect a tumor even if it is embedded within the glands tissues of the breast.

REFERENCES

EXPERIMENTAL ASSESSMENT OF MICROWAVE DIAGNOSTIC TOOL FOR ULTRA-WIDEBAND BREAST CANCER DETECTION

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Abstract—An ultra-wideband microwave imaging system that employs a heterogeneous breast phantom and covers the ultra-wideband (UWB) frequency range (3.1 GHz to 10.6 GHz) is presented. The platform scanning system allows monostatic and bistatic mode of operation. In this work, developed heterogeneous phantoms are used to mimic the realistic breast tissues. A utilized tapered slot antenna array allows for a high resolution hemispherical scan, achieved by rotating the imaged object on a turntable. Full design details of the scanning system and the utilized post-processing algorithm are explained. To validate the reliability of the presented system, the results of several imaging cases, including the challenging low dielectric contrast case, are presented.

1. INTRODUCTION

At microwave frequencies, breast imaging methods have been explored for several decades. There are three different approaches to microwave imaging, namely the passive, hybrid and active methods. In the passive method, the tumor is detected based on the increase in its temperature compared to normal tissues. Hybrid methods use microwave energy to rapidly heat tumours and using ultrasound transducer to detect pressure waves generated by the elasticity properties of the heated tissues. Active methods involve illuminating the breast with microwaves and measuring the scattered signals.
The use of microwave imaging for biomedical applications was initiated by Jacobi et al. in the late 70s. Antennas immersed in the water were designed to obtain images of internal structure of a canine kidney [1]. During the last decade, research activity in microwave imaging has mainly focused on the breast cancer detection.

In 2003, a medical model where microwave breast imaging is performed through a water-coupled boundary to detect breast cancer was developed by the school of engineering at Dartmouth College, USA [2]. In this experiment, modulated, continuous wave signals were transmitted from sixteen-monopole antennas operating over a frequency between 500 MHz to 3 GHz. Nine antennas at the other half of the imaging array (glycerin mixture) are used to receive the scattered signals. A superheterodyne technique is used to extract the phase and amplitude of the high frequency reflected signals. The reflected microwave signals are then converted into a 2D model map relating to the dielectric permittivity of breast tissues. However, the antenna array has to be moved manually via a hydraulic jack for the collection of multiple planar data required to get a model map of the breast tissues.

In 2009, Klemm et al. from the University of Bristol employed the UWB microwave radar using a real aperture array of UWB antennas that operate in a multi-static mode [3]. Antennas are positioned on a section of the hemi-sphere, conforming to the curved breast shape. The array is formed around the lower part of a 78 mm-radius sphere, in four rows of four antennas [4]. All antennas are aligned in rows and columns, and the array has two axes of symmetry.

In this paper, an automated high resolution hemispherical imaging system using ultra-wideband microwave signals is reported. The presented system enables a cylindrical scanning of a heterogeneous breast phantom. The main contribution of this paper is the integrated imaging system that includes a realistic breast phantom, directive antenna, and scanning platform. That system besides the presented imaging algorithm forms a complete diagnostic tool for breast cancer detection. To validate the reliability of the system, imaging results for three cases of early tumors are presented.

2. FABRICATION OF HETEROGENEOUS BREAST PHANTOMS

Homogenous phantoms have been of considerable use to test the principles of microwave imaging [5–7]. However, they are insufficient to test the feasibility of UWB imaging with respect to real breast tissues because of the composition of adipose and fibro glandular tissue [8].
Breast tissues heterogeneity is quite complex and different from one person to another. The phantom fabrication is not only motivated by the need for realistic breast phantom that can mimic the geometry and the dielectric properties of human breast, but also by the need for set of phantoms that can represent the range of human breast densities. Breast density is used to refer to the percentage of adipose and fibro-glandular in the breast. In the developed phantom, it refers to the percentage of high dielectric material and low dielectric materials that represent adipose and fibro-glandular, respectively.

The developed phantom emulates the dielectric properties of human breast over the UWB frequency range of 3.1 GHz to 10.6 GHz [9, 10] and covers a greater range of dielectric properties of normal tissue. Data published by Campbell and Land [11] reported on the high contrast between normal (fat and other tissue) and cancerous tissue. Another significant finding was that the heterogeneity and the complexity of the breast are higher than the previously reported. This is due to the almost random locations of the fat tissues with a low dielectric constant and the fibro glandular tissues with high dielectric constant. Heterogeneity of normal breast tissue has been underestimated by many early researchers [12, 13]. In 2007, a comprehensive study has suggested that the location of which the normal tissue samples were taken was the reason the heterogeneity was overlooked by previous studies [14]. The high dielectric constant of gland makes the contrast of normal tissues to cancerous tissues small, and thus, creates a challenge in the microwave detection methods.

The two breast phantoms used in our system, namely phantom A and phantom B, are fabricated to represent highly dense and low dense breasts as categorized in breast classification done on real breasts from reduction surgeries [15]. A low density breast is considered to have high fat content and low fibro glandular composition in the breast. Conversely, the high density breast has low fat content but high fibro glandular composition. In reality, the breasts of younger women contain less fat and are denser than older women’s breasts. Thus, it is possible to assume that the developed phantom A represents breast of young women, whereas phantom B represents breast of older women. It is worth mentioning that a study [16] reveals that the risk of getting breast cancer increases with age.

Agar based material has been chosen in the fabrication of our phantoms since it can be easily shaped and has a stable dielectric properties over a long period of time. A pyrex glass mould is used to give a hemispherical shape to the phantoms. The hemispherical phantoms used in the experiment have a diameter of 13.4 cm. A random composition of fibro glandular mixture is deposited into the
base material of the breast phantom (fat tissue material) to simulate the heterogeneity [17]. By varying the composition of low dielectric materials such as cornflour and water, the dielectric properties of the phantom can be easily altered. In particular, the proposed phantom can represent varieties of women breasts from a low dense breast (dominance of adipose tissue), to a high dense breast (more heterogeneous mixture of fibro glandular and adipose). For the preparation of the tumor phantom, 7.68 g grape seed oil and 0.71 g detergent was mixed in a beaker. In a separate beaker, formaldehyde solution (0.313 g, 32%) and $p$-toluic acid were mixed by shaking with 3.14 g 1-propanol. 75 g milli-Q water was heated and the oil-detergent mixture was added. The formaldehyde and $p$-toluic acid mixture were then added together with 13.56 g of agar in small portions. 0.1 g alizarin dye was added to colour the tumor phantom which gives a red colour. The mixture was placed in suitable moulds and allowed to cool to room temperature [9].

The percentage of materials used in the developments of the two phantoms, in addition to the average and range of values for the dielectric constant, are shown in Table 1. Phantom A has a higher mean dielectric permittivity of 31.7 than phantom B that has a lower mean value of 28. The high volume percentage of agar mixture (high dielectric) with 60% composition and only 40% of corn flour mixture (low dielectric) making it have a high range of dielectric permittivity of 16.66 because the high concentration of dielectric constant of agar mixture which was randomly distributed in the phantom A. This value

![Figure 1](image)

**Figure 1.** Measured dielectric constant of breast phantom samples for low and highly dense breast.
Figure 2. Measured conductivity of breast phantom samples for low and highly dense breast.

Table 1. The contents of phantoms (A and B), and mean and range of dielectric permittivity.

<table>
<thead>
<tr>
<th>Phantom</th>
<th>% Volume of corn flour and water</th>
<th>% Volume of Agar mixture</th>
<th>Mean Dielectric Permittivity</th>
<th>Range of dielectric permittivity</th>
</tr>
</thead>
<tbody>
<tr>
<td>A (Dense Breast)</td>
<td>60</td>
<td>40</td>
<td>31.7</td>
<td>16.66</td>
</tr>
<tr>
<td>B (Low Dense Breast)</td>
<td>80</td>
<td>20</td>
<td>28.0</td>
<td>11.72</td>
</tr>
</tbody>
</table>

confirms the high variability of the high density phantom A due to the extreme composition of low and high dielectric mixtures. The measured dielectric properties of the breast phantoms are depicted in Figure 1 and Figure 2. In this result, each of the curves with vertical bars represents the average dielectric properties of the breast phantom A and phantom B. The vertical bars also show the variability or the standard deviation of the measured dielectric properties of the phantoms from the average value. The heterogeneous hemispherical phantoms breast phantom used in the system has an average dielectric properties ($\varepsilon = 18$ to 40 and $\sigma = 0.9$ to 7 S/m) to simulate normal breast tissues across the UWB frequency range. For the target (tumor), a mixture of agar, oil and formaldehyde solution is used with properties of ($\varepsilon = 52$ and $\sigma = 1.8$ S/m).
Figure 3. The bottom (a) and top (b) views of the tapered slot antenna.

3. SCANNING SYSTEM

3.1. Antenna

In the designed imaging system, a directive tapered slot antenna was used [18]. It is fed by a tapered microstrip-line with a suitable microstrip-slot transition. The slot is gradually tapered along the $x$-axis and symmetric along the $y$-axis towards the feed. The tapered slot is defined by an exponential function that is optimized for the best possible performance. The microstrip-slot transition, needed to achieve strong coupling between the microstrip feeder and tapered slot radiator, uses virtual open and short circuits in the form of a radial slot stub and a radial microstrip stub. From parametric simulations, it was found that by removing ellipse shaped regions from the conductive regions, the antenna produces a higher directivity in the lower part of the UWB [18]. The antenna is fabricated on Rogers RT6010LM substrate, featuring a dielectric constant of 10.2, a loss tangent of 0.0023, and thickness of 0.64 mm. The developed antenna (Figure 3) has a compact size of $36 \, \text{mm} \times 36 \, \text{mm}$. The optimum dimensions of the tapered slot for an ultra-wideband performance are found to be $H = 36 \, \text{mm}$, and $L = 30 \, \text{mm}$. As can be seen from Figure 4, the antenna operates from 3.1 GHz to over 10.6 GHz for the 10 dB return loss reference.

3.2. Scanning Platform

The scanning system configuration of the proposed UWB microwave imaging system is shown in Figure 5. The system uses a USB
Figure 4. The reflection coefficient at the input port of the antenna.

Figure 5. Scanning platform.

interface to activate a stepper motor to rotate the phantom at a minimal angular step of 0.72°. The developed system supports monostatic and bistatic mode of operation. In the work presented here, however, monostatic operation was used where a single antenna element is used for transmitting and receiving of the UWB pulses. The UWB pulses are generated using ZVA24 Rohde and Schwartz vector network analyser (VNA) in a step-frequency manner typically using 401 equidistant frequency points across the UWB. Obtaining the complex $S$-parameters is done using the virtual instrument software architecture (VISA). This is a standard for configuring and programming instrumentation via a variety of buses such as GPIB,
RS232, Ethernet and USB. The presented system employs a common Ethernet interface and requires less than 3 seconds to obtain 401 discrete complex numbers. In this system, a set of the ultra-wideband antenna elements depicted in Figure 3 can be offset from each other in the vertical and azimuth directions to minimize adverse effects of mutual coupling. Assuming that the array is limited to four antenna elements, ultra-wideband (UWB) data for image reconstruction can be acquired by antennas connected to a 4-port vector network analyser. To achieve a hemispherical scan, the imaged object is placed on the turntable (rotational axis) for a 360° scan while the elements of the array antenna are fixed and suitably spaced in vertical and azimuth directions. An adjustable antenna holder varies a radial position of the antenna from the phantom, depending on the size of the antenna and the phantom and also the antenna height.

4. RESULTS AND DISCUSSIONS

The designed system is used to image the fabricated heterogeneous phantoms. 50 antenna positions are considered with 7.2 angular steps. The antenna was placed at different horizontal planes. However, the best images that show the target clearly are obtained when the antenna is located at the same horizontal plane as the emulated tumor.

For the post-processing procedure needed to produce images and to quantify them, the algorithm and metric formulas proposed in [19] are adapted for this work. The underlying method of the algorithm is based on using transmitted ultra-wideband signals and recording of the time-domain back-scattered signals. This data has echo signals which ideally originated from electromagnetic scattering targets (e.g. breast tumor). The algorithm works by making a hypothesis that an echo signal originated from a given point; the normalized difference signals of each antenna are added at this space location. If the hypothesis for the particular scatterer location is correct, the signals add coherently and a large value of the sum is obtained. If the hypothesis is incorrect, the signals add incoherently and the sum is small. A continuous colour image is produced using a shading operator to interpolate at non-tested points. Strong intensity colours indicate the location of significant scattering objects.

Quantitative metrics are used to evaluate the produced image. In order to define the metrics used in this paper, it is necessary to first define several objects: $p$ denotes any $(x, y)$ point inside the body to be imaged (phantom); $Z$ is the set of all discrete points; $T$ is the set of $(x, y)$ points that map to the location of the emulated tumor in the phantom. The function $I(p)$ gives the image intensity at point $p$. The
Figure 6. Imaging results of (a) Phantom A, (b) Phantom B, and (c) Phantom A with two targets inserted.

The first metric used is the ratio of the average intensity value of points located in the tumor region over the other points located in normal breast tissue. It is given as

$$Q = \frac{\mu[I(p)]}{\mu[I(p)]} \forall p \in T \forall p \notin T$$

(1)

where $\mu[\cdot]$ denotes the mean function. A higher value for this metric implies the tumor intensity is more intensive than the background regions.

The second metric $\gamma$ is the ratio of the maximum intensity value of the tumor region over the maximum intensity of the complete image [20]. It is given as

$$\gamma = \frac{\max[I(p)]}{\max[I(p)]} \forall p \in T \forall p \notin Z$$

(2)

where the function max[·] returns the maximum image intensity of the specified set of points. If this value is 1, it implies that the tumor is the strongest scatterer. If this value is less than 1, it quantifies the significance of the tumor in terms of electromagnetic scattering.

The above algorithm was used for post-processing. The imaging results of the phantoms are shown in Figure 6 for dense and low dense...
Figure 7. Imaging result when there is no target inserted.

phantoms. The low intensity color bar represents the low dielectric material (fat tissue), and high intensity color bar represents the high dielectric materials (fibro glandular and tumor tissue). The diameter of the target is 0.6 cm. The actual positions of the embedded targets in the utilized phantoms are indicated by the black circles in Figure 6. For phantom A with two targets, target T1 has 0.5 cm radius of a closed end cylinder filled up with water, whereas target T2 is the tumor fabricated material.

It is clear from the presented results in Figure 6 that the designed system as whole is able to detect small tumors even in a highly dense breast with a dielectric contrast as low as 1:1.3. The use of the system for the detection of two tumors in a dense breast was also successful as revealed in Figure 6(c) although there is a slight shift of a few millimeters in the central position of the detected targets in comparison with their actual positions. To verify that the system does not show false targets, the breast phantoms were imaged without an emulated tumor as shown in Figure 7 where no target can be seen.

To quantify the success of the imaging system, the metrics parameters are calculated for the imaging results. Table 2 shows the $Q$ and $\gamma$ metric results for the given images in Figure 6. Phantoms A and B show the image intensity at the tumor region was 1.9 and 2 times, respectively, more intense then the background intensity. The $\gamma$ metric in both cases was 1 indicating the tumor was the strongest electromagnetic scatterer in both cases. For the phantom with two targets, which are denoted $T_1$ and $T_2$ as shown in Figure 6(c), the $\gamma$ value was 1 for $T_2$ while $T_1$ was slightly lower at 0.96. This suggest that $T_1$, consisting of water, is a slightly better electromagnetic scatterer than $T_2$ which consisted of material mimicking a tumor. Furthermore, the $Q$ values for both targets were 1.795 and 1.73 respectively again showing the tumor intensity is at least 1.73 times higher than the background intensity.
Table 2. Performance of the algorithm for the phantoms with single and two targets.

<table>
<thead>
<tr>
<th>Phantom</th>
<th>Q</th>
<th>γ</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>1.9</td>
<td>1</td>
</tr>
<tr>
<td>B</td>
<td>2</td>
<td>1</td>
</tr>
<tr>
<td>Two Targets</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>$T_1$</td>
<td>1.795</td>
</tr>
<tr>
<td></td>
<td>$T_2$</td>
<td>1.73</td>
</tr>
</tbody>
</table>

5. CONCLUSION

This article reports the development of an ultra-wideband microwave imaging system using tapered slot antennas and heterogeneous breast phantoms aimed for breast cancer detection. The fabricated breast phantom closely mimics the realistic breast electrical properties and emulates the heterogeneity of real breast. Two classes of breast phantoms representing different breast density classifications are used in the imaging system. A reconstruction algorithm based on confocal imaging is used for post-processing. The obtained results indicate the possibility of detecting small tumors in situations with as low as $1:1.3$ contrast in the dielectric constant. A future work will include the imaging of the phantom in 3D.

REFERENCES


COMPACT WIDEBAND ANTENNA IMMERSSED IN OPTIMUM COUPLING LIQUID FOR MICROWAVE IMAGING OF BRAIN STROKE

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Abstract—This article reports on the design of a wideband compact microstrip-fed tapered slot antenna aimed at microwave imaging of a brain stroke. The antenna is immersed in a carefully designed coupling liquid that is used to facilitate higher signal penetration in the brain and thus increased dynamic range of the imaging system. A parametric analysis is used to find out the required properties of the coupling liquid. A suitable mixture of materials is then used to implement those properties. In order to protect the antenna from the adverse effects of the coupling medium, dielectric sheets are used to cover the radiator and the ground plane. To verify the proposed design in brain imaging, the antenna is tested using a suitable head model. It is shown that the antenna with a compact size (24 mm × 24 mm) on RT6010 substrate (dielectric constant = 10.2) operates efficiently over the band from 1 GHz to more than 4 GHz with more than 10 dB return loss. The time domain performance of the antenna supports its capability to transmit a distortion-less pulse with a high fidelity factor inside the head tissues.

1. INTRODUCTION

Microwave imaging systems for medical diagnostics have recently been proposed to augment conventional medical imaging systems [1]. Research has shown that there can be significant differences in dielectric properties of normal and abnormal tissues at microwave frequencies [2–6]. This phenomenon is well known in the field of microwave imaging (MI) for breast cancer detection. However, MI
techniques applied to other tissue abnormalities, stroke detection in particular, are yet to be adequately researched. Potentially, MI offers a diagnostic method with a wide variety of advantages, such as non-ionizing radiation, low-cost portable system and fast imaging results.

A stroke is an abrupt onset injury that affects the central nervous system. It can be classified into two main categories ischemic and hemorrhagic stroke. Ischemic stroke results from blocking the artery that carries blood to the brain. Hemorrhagic stroke results from bleeding within the brain or in the space surrounding. Both medical conditions lead to death in the intermediate future if left untreated. Moreover, the symptoms can be similar between the two conditions, however, the medical treatment is significantly different. An incorrect determination of the type of stroke most certainly leads to the death of the patient.

A clinical decision has to be made within 3 hours of the onset of the symptoms of a stroke to ensure the treatment is effective [7]. It is for these reasons MI systems for stroke detection has gained significant research interest in recent times; examples are found in [3] and [5].

To achieve sufficient penetration in the head tissues, it is believed that the microwave imaging system must operate at approximately 1–4 GHz [3–5]. Thus, a wideband antenna that operates efficiently across that band is a crucial element in the success of microwave imaging system of the brain.

This paper reports the design of a miniaturized tapered slot antenna with an operational band that extends from 1GHz to 4GHz. The performance of the antenna is tested via simulations and measurements. The SAM (Specific Anthropomorphic Mannequin) head model [8] is used in the simulations to test the proposed antenna. Both the antenna and the head phantom are immersed in a coupling liquid to improve the signal penetration. Since the post-processing of scattered microwave signals in brain imaging generally requires time-domain signals, the fidelity factor of the transmitted time-domain pulses from the antenna into the coupling medium is examined.

2. ANTENNA AND COUPLING MEDIUM DESIGN

Different types of antennas have been previously designed to operate in imaging systems [9–11]. In this paper, the utilized antenna is aimed to be compact in size, possess directional properties and provides resonance across the desired frequency band. Fig. 1 shows the configuration of proposed tapered slot antenna which is designed for a microwave-based brain imaging system. It is fed using a microstrip line of 50Ω characteristic impedance. The slot of the antenna is tapered
Figure 1. Configuration of the designed antenna (a) without cover, (b) with protecting cover.

using an elliptical function [12, 13]. The antenna is designed using Rogers RT6010 (thickness = 0.635 mm, relative dielectric constant = 10.2) as the substrate. Giving the lowest frequency of operation \( f_1 \) and the dielectric constant \( \varepsilon_r \), the width and the length of antenna designed using the following equation:

\[
W = L = \frac{c}{f_1} = \sqrt{\frac{2}{\varepsilon_r + 1}}
\]  

\( c \): is the speed of the light.

The radiating structure of the antenna is the intersection of the quarter of two ellipses with major radii (\( r_1 \) and \( r_2 \)) and secondary radii (\( r_{s1} \) and \( r_{s2} \)) using the following equations:

\[
r_1 = \frac{W}{2}
\]  

\[
r_2 = \frac{W}{2} - wf
\]  

\[
r_{s1} = L - a
\]  

\[
r_{s2} = 0.5r1
\]

\( a \): is a parameter used to control the frequency of operation.

In the current design \( r_1 = 12 \text{ mm}, r_2 = 11.55 \text{ mm}, r_{s1} = 23.72 \text{ mm}, r_{s2} = 6 \text{ mm}, a = 0.28.\)
Next, a miniaturization technique is used to reduce the size of the structure [14, 15]. The modification from the traditional tapered slot structure starts by removing the tapered ground and the slotline-to-microstrip transition. A direct connection is used to connect the top radiator with the microstrip line, whereas the bottom layer is used as ground plane. The slot $s$ between the top radiator and the ground is used to achieve fine quality matching with the feed point. The feed structure is curved from the edge of the structure to ease the connection of the microstrip line with the external port in the direction that is suitable for the planned imaging system.

A symmetrical corrugation is then used in the outer edges of both the top radiator and the background in order to miniaturize the size of the antenna. Those corrugations are used in order to increase the effective path length of the surface current, and thus, to enable the acceptable performance of the antenna at the low part of the band when the size of the antenna is reduced. It can also help to suppress standing waves arising in the antenna’s structure [14].

To reduce the scattered signals at the interface between the antenna and the head tissues, and to achieve the best matching with the tissues, both the antenna and the imaged object are immersed in a coupling medium with a high dielectric constant and low conductive loss [16, 17]. The high permittivity of the liquid can physically reduce the size of the antenna and potentially increase the dynamic range of the imaging system. In order to design a proper coupling liquid, different mixtures were tested. HP85070B coaxial probe is connected to the HP network analyzer (HP8530A) and used to measure the dielectric properties of different kinds of materials. It is found that a mixture of 70% water and 30% of solution that includes the same

![Figure 2. Variation of permittivity and conductivity with frequency for the designed coupling liquid.](image-url)
percentage of grape seed oil and polysorbate 80 (Tween-80) achieves the best possible matching between the designed antenna and the utilized head phantom. The measured variation of the dielectric permittivity and conductivity of the designed liquid for the frequency range from 1–4 GHz is shown in Fig. 2. In order to optimize the design of the proposed antenna when immersed in the designed coupling liquid, the properties of the manufactured coupling medium is loaded into the simulation tool. The antenna is also tested when operating in front of a realistic SAM head model as depicted in Fig. 3.

In order to protect the antenna from the adverse effects of the coupling liquid, the top radiator and the ground plane are to be covered by a suitable protective material, such as resin or varnish. In the current design, it is covered by a dielectric sheet that has the same dielectric properties of the substrate as shown in Fig. 1(b). An adhesive material that has a dielectric constant close to that of the utilized substrate is used to glue the antenna and the covering sheets together. Given the lowest frequency, the thickness and the dielectric properties of the substrate, the antenna was initially designed according to the guidelines presented in [12]. The length of the slots of the corrugated structure is chosen to be quarter of the effective wavelength calculated at the center frequency of operation. Since the antenna is designed to operate across the band from 1 GHz to 4 GHz, the center frequency is 2.5 GHz. The dimensions of the antenna and the slots are then optimized using CST Microwave Studio. The final dimensions in (mm) are: $S = 0.28$, $w_f = 0.45$, $L_s = 3.5$, $W_s = 0.5$, $d_s = 0.5$, $w = 24$, $L = 24$, and $W_{ed} = 17$. 

Figure 3. The antenna when used to image the SAM head phantom.
3. RESULTS AND DISCUSSION

Performance of the proposed antenna is first verified via computer simulations. Next, the antenna is manufactured (Figs. 4(a) and (b)) and tested to confirm its simulated performance. The antenna was tested while immersed in the designed coupling liquid (Fig. 4(c)) with and without the presence of a head phantom.

The simulated and measured return loss of the antenna when immersed in the coupling liquid and without the head phantom is shown in Fig. 5. The obtained results indicate that the antenna has a

Figure 4. Photo of manufactured antenna. (a) Top radiator and (b) ground side without and with cover. (c) The test platform with two antennas immersed in the coupling liquid.
reflection coefficient of less than $-10\,\text{dB}$ across the required band from 1 GHz to 4 GHz. The plastic container that includes the antenna and the coupling medium has a slight effect on the performance as indicated in Fig. 5(a). To test the effect of the corrugations on the covered band, the reflection coefficient was calculated using the simulation tool for the antenna without corrugations. It is found that the reflection coefficient in that case is less than $-10\,\text{dB}$ across the limited band from 2.35 GHz to 4 GHz as shown in Fig. 5(a).

To test the directive properties of the antenna, the gain was calculated using the simulation tool. It was found that the gain varies between 4 dBi and 6.7 dBi across the band from 1–4 GHz. Without corrugations, the gain changes between 3.5 dBi and 5.5 dBi across the same band as depicted in Fig. 5(b).

To verify the effectiveness of the chosen coupling liquid that has an average dielectric constant of around 43 as depicted in Fig. 2, the reflection coefficient of the antenna is calculated via simulations for
Figure 5. (a) The measured and simulated reflection coefficient of the antenna when immersed in the designed coupling liquid in the absence of the head phantom, (b) gain of the antenna with and without corrugations, and (c) the simulated reflection coefficient for other coupling liquids.

Figure 6. The simulated reflection coefficient of the antenna placed in front of SAM head model and immersed in different coupling liquids with the shown average dielectric constant.

three different types of coupling liquids that have an average dielectric constant of 20, 30, and 40. The result of simulations shown in Fig. 5(c) reveals that the liquid with an average dielectric constant of 40, which is close to that of the fabricated one, enables the antenna to achieve its best performance.

The simulated return loss of the antenna when in close proximity
Figure 7. The time domain response of the manufactured antenna.

Recent reported brain imaging systems uses time-domain pulses to reconstruct the image (see for example [5]). For this reason, the time domain impulse response of the antenna is tested to verify its capability to support the transmission/reception of narrow pulses in a distortion-less manner. Two antennas are placed at the same height above the ground with 3 cm distance between them. The two antennas face each other in the end-fire direction. The space between the antennas is filled with the developed coupling liquid. The vector network (R&S ZVA24) is used to generate a narrow pulse with 1–4 GHz frequency content. The pulse transmitted from one of the antennas is received by the other antenna. The measurements are shown in Fig. 7. It is clear the developed antenna supports almost distortion-less transmission which minimises so-called ghost targets occurring in the microwave imaging system. It is worth mentioning that the time delay between the transmitted and received pulses in Fig. 7 is due to the antennas’ structure, connecting cables and the 3 cm distance between the end points of the two face-to-face antennas.

To quantify the distortion level in the transmitted pulses inside the coupling liquid, the second antenna is moved away from the first one with different distances so that the variation of the fidelity factor as a function of distance inside the coupling medium can be calculated. The fidelity factor is calculated as the maximum magnitude of the cross correlation between the observed pulse at a certain distance and the excitation pulse [18]. The simulated and measured results are shown in Fig. 8. The results indicate that the fidelity factor of the antenna decreases as the distance from the antenna increases. However, the
value is still within the acceptable limit for a successful imaging system (above 50%) as compared with previous reported values [19]. There are some differences between the simulated and measured fidelity factor as depicted in Fig. 8. The effect of the container on the performance of the antennas can be predicted from the simulated results with and without the container in Fig. 8. It is clear that the multiple reflections from the boundaries of the plastic container cause a slight degradation in the fidelity factor.

In order to clarify the importance of using the designed coupling liquid, the fidelity factor is also estimated inside the SAM head model for two cases. First, when the head and antenna are immersed in the developed coupling liquid and the second one when the antenna

![Figure 8](image)

**Figure 8.** The simulated and measured fidelity factor as a function of distance from the antenna when it is immersed in a coupling liquid.

![Figure 9](image)

**Figure 9.** The simulated fidelity factor as a function of distance from the antenna to the SAM head model with and without coupling liquid.
is redesigned to work in free space (without coupling liquid) with dimensions of $(99.5\,\text{mm} \times 99.5\,\text{mm})$. The results depicted in Fig. 9 show that the fidelity factor of the antenna immersed in the coupling liquid is better than its value when no coupling liquid is used despite the need for a larger antenna size in the no-coupling liquid case. For the antenna presented in this paper, the fidelity factor is within reasonable values [19] inside the head phantom when it is immersed in the designed coupling liquid.

From comparing the results of the fidelity factor with (Fig. 9) and without (Fig. 8) the presence of the head phantom, it is clear that the presence of the phantom causes a noticeable reduction in the fidelity factor. This is one of the challenges facing the design of a reliable microwave-based brain imaging system.

4. CONCLUSION

The design of a compact wideband tapered slot antenna immersed in a coupling liquid for the use in a microwave-based brain imaging is presented. The coupling liquid is designed properly to improve the matching between the antenna and the brain tissues using a mixture of water, grape seed oil and polysorbate 80 (Tween-80). To miniaturize the antenna, corrugations are introduced in outer edges of both the radiator and the ground plane. To protect the antenna from ill effects, such as corrosion of the conductive layers by the coupling liquid, the antenna is covered by a dielectric sheet. The simulated and measured return loss of the antenna tested with and without a head phantom have shown that the antenna covers the band from $1\,\text{GHz}$ to $4\,\text{GHz}$ with better than $10\,\text{dB}$ return loss across the desired band.

The time domain performance of the antenna has also been studied. It has been shown that, although the fidelity factor decreases as the signal penetrates the head, the value of that factor is still within the acceptable limits when the coupling liquid is used.

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Single Layer Reflectarray With Circular Rings and Open-Circuited Stubs for Wideband Operation

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Abstract—The design of a single-layer reflectarray, which employs a new phasing element in the form of a fixed-size circular ring and a variable-length open-circuited stub, is presented. The array is developed on a thin substrate supported by a thick foam material. Investigations are performed to obtain a linear reflection phase as a function of the stub’s length when the element operates in a unit cell. This goal is achieved by a suitable choice of the ring’s radius and width and the stub’s width. In order to validate the simulated element’s reflection phase behavior, a waveguide simulator is manufactured to perform experimental tests. The phasing element offering best linear phase characteristics is used to design an X-band offset fed 13 x 13 element reflectarray pointing at 20° from the broadside direction. Full-wave simulations performed using CST Microwave Studio show desired radiation characteristics of the designed array antenna. The simulated performance is confirmed by experimental tests performed on the fabricated reflectarray prototype showing a 17.8% 3-dB gain drop bandwidth.

Index Terms—Microstrip antennas, phase shifters, reflectarrays.

I. INTRODUCTION

Microstrip reflectarray antenna formed by a planar array of microstrip patch elements is an attractive alternative to a curved reflector antenna and a conventional phased array [1]. It uses variable size microstrip patch elements as phase shifters to convert a spherical wave incident from a feed horn into a plane wave in a specified direction. Its advantage over a conventional phased array is that it uses a low-loss spatial beamforming network [2], [3]. Power losses of such type of beamformer are frequency independent and are smaller than those of a circuit type beamformer, especially for large arrays operating at upper microwave or millimeter wave frequencies. When compared with a curved reflector, it is easier to transport and deploy because of its planar reflector structure. However, its disadvantage is a narrow operational bandwidth (only a few percentages) when variable size microstrip patches are used. To counter this drawback, microstrip phasing elements having a reflection phase range not less than 360° with a slow phase slope as a function of their size are required.

Multilayer stacked patches of variable size were proposed as phasing elements with multiples of 360° phasing range and reduced phase slope as a function of dimensions. The proper choice of the dimensions also leads to approximately linear phase characteristics and this is welcome from the manufacturing point of view as the phase errors due to manufacturing tolerances are reduced. Wideband operation of such multilayer microstrip reflectarrays was demonstrated [4]. However, the multilayer solution is offset by a labor-intensive manufacturing process. To counter that, a renewed interest in single layer reflectarrays has been observed in recent years.

It has been demonstrated that by using variable size multiresonance planar elements in the form of circular, elliptical or rectangular rings supported by thick foam materials, an improved reflection phase range (exceeding 360°) accompanied by a slow phase slope can be achieved [5]-[7]. However, obtaining a linear phase from these elements is challenging from the manufacturing perspective due to the need for very narrow gaps between rings. To overcome this problem, the work presented in this paper focuses on using a new planar phasing element formed by a fixed-size circular ring accompanied by a variable length arc.

The arc, which is connected to the ring, forms a variable length open-circuited stub for the reflectarray phasing. Using this arrangement, the phase range can exceed 360°. An initial idea of using this phasing element was put forward in [8] without paying attention to the choice of the ring’s radius and width, or the arc’s width. As a result, a highly nonlinear reflection phase as a function of arc’s length was generated [8]. Further work to obtain a liner phasing, but limited only to simulations, was presented in [9].

In this paper, full wave simulations using CST Microwave Studio are performed to obtain linear phase characteristics of this type of reflectarray phasing element. It is shown that the required phasing characteristics can be achieved using the properly chosen radius and width of the circular ring and the width of the variable length arc attached to the ring. The investigations are performed in a unit cell environment. In order to experimentally validate the simulation results, a waveguide simulator is developed. The phasing element showing the most linear phase behavior as a function of the arc length is used to form an offset-fed 13 x 13 element reflectarray for operation at the X-band with the center frequency of 11.5 GHz. The performance of this array is investigated by full wave simulations and experiments.
II. PHASING ELEMENT DESIGN

A. Unit Cell Configuration

Investigations focus on a single layer reflectarray operating in X-band with the center frequency of 11.5 GHz. The proposed phasing element is depicted in Fig. 1(a). It is composed of a circular ring of width $\Delta R$ and radius $R_1$ with a circular arc of radius $R_2$ attached to it using a strip of height $H$. The electrical length of the arc $\theta_0$ is the main parameter responsible for controlling the phase of reflection coefficient when a wave is incident on an array of such antenna elements. It is supported by a thin substrate of thickness 0.508 mm and relative permittivity 2.2 and by a thick foam of thickness $T_2 = 3.175$ and relative permittivity $\varepsilon_2 = 1.06$, as shown in Fig. 1(b).

The role of foam placed under the substrate is to reduce the slope of the reflection phase characteristic so that the operational bandwidth can be increased and manufacturing errors minimized. However, an introduction of such material reduces the phase range and thus it is important to check that the required 360° phase range is still maintained.

B. Phasing Element Analysis and Optimization

The analysis concerns the proper choice of the radius and width of the ring and width of the arc operating as an open-circuited stub. The aim is to obtain a linear reflection phase characteristic so that the operational bandwidth can be increased and manufacturing errors minimized. However, an introduction of such material reduces the phase range and thus it is important to check that the required 360° phase range is still maintained.

The proposed two-port representation is valid under the assumption that the stub weakly interacts with the vertically polarized incident wave. In this case, the input reflection coefficient at Port 1 can be represented as a bilinear function of the reflection coefficient of the variable length stub [11]

$$\Gamma_{IN} = S_{11} + \frac{S_{12} S_{21} \Gamma_{Stub}}{1 - S_{22} \Gamma_{Stub}}$$

(1)

where $\{S_{ij}\}$ are the $S$-parameters of the equivalent two-port and $\Gamma_{Stub} \approx e^{i2\beta L}$, $\beta$ is the phase constant and $L$ is the length of the stub.

The linear transformation $\Gamma_{IN} = S_{11} + S_{12} S_{21} \Gamma_{Stub}$ can be obtained from (1) when $S_{22} \approx 0$. This occurs when the input impedance looking at Port 2 is equal to the characteristic impedance of the stub.

The above explanations indicate that if an approximately linear behavior of $\Gamma_{IN}$ as a function of $\Gamma_{Stub}$ is the design objective, the radius and width of the ring as well as the width of the stub/arc have to be suitably chosen.

In order to find the parameters offering a linear reflection phase curve, the phasing element is divided into two parts, the ring including the vertical feed, and the arc. A discrete port is assigned at the feed point of the ring, and then an input impedance of the ring as a function of the ring width $(\Delta R_1)$ at the center frequency of 11.5 GHz is investigated. Assuming the ring radius $R_1 = 2.65 \text{ mm}$, the input impedance of the circular ring as a function of the ring width $(\Delta R_1)$ is plotted in Fig. 3(a) assuming a square unit cell of $\Omega_x = \Omega_y = 15 \text{ mm}$, which is equivalent to 0.58 wavelength $(\lambda)$ at the design frequency of 11.5 GHz.

As observed in Fig. 3(a), the ring input resistance is inversely proportional to the ring width and is approximately equal to 150 $\Omega$ when $\Delta R_1$ is around 0.09 mm. Fig. 3(b) shows the characteristic impedance of a straight microstrip line as a function of the strip width $W$ for a composite dielectric formed by the substrate and the foam. The microstrip characteristic has an inverse relationship with its width. In order to have the characteristic impedance of the microstrip arc utilized in the presented design at around 150 $\Omega$, its width $W$ has to be 2 mm.

Eventually, the final dimensions of the phasing element are obtained by generating a family of reflection coefficient phase curves for a unit cell at 11.5 GHz when the ring’s and arc’s width are varied. Table I shows the final optimized parameters that produce the best linear reflection phase curve for the unit cell at 11.5 GHz.

C. Characteristic Results

The phase characteristics are produced by performing fullwave electromagnetic simulations of a unit cell using Frequency Solver of CST Microwave Studio 2010. The simulations assume a normal wave incidence on the unit cell. In this case, the unit cell operation is equivalent to that of a TEM waveguide accommodating the phasing element [10].

The reflection phase as a function of the electrical length of the phasing arc $(\theta_0)$ at 11.5 GHz is illustrated in Fig. 4.

The presented results show that the reflection phase has a sufficient phasing range of around 410°. It has a linear variation in
The phasing characteristics for lower (11 GHz), center (11.5 GHz), and upper (12 GHz) frequencies are compared in Fig. 5. The phase range at each of the chosen frequency exceeds the required 360° and the phase slope stays approximately constant.

These phasing characteristics are the basis for obtaining an increased operational bandwidth of a reflectarray formed by the proposed phasing elements.

In order to show whether the phase performance of the proposed element is critical to a slight change in the angle of incidence, the phase is simulated for different angles. The results presented in Fig. 6 show the phase performance of one cell for oblique incidence in both the principal plane (\(\varphi = 0^\circ\)) and off principal plane (\(\varphi = 30^\circ\)). It can be clearly seen that the variation in the performance between normal and oblique incidence of up to 30° is small in both the principal and off principal planes. For more accurate design, those changes, albeit being small, can be included as part of the design for different cells of the reflectarray.
D. Validation by Waveguide Simulator

In order to confirm the validity of simulation results for the reflection phase characteristics, an X-band waveguide simulator [12]–[14] is developed. Because the waveguide uses all conducting walls, the condition of TEM wave used in the earlier undertaken simulations, when producing phase curves, cannot be met. In the waveguide simulator, the incident angle \( \theta \) of the inbound wave is dependent on an operating wavelength and cut-off wavelength of the waveguide, as given by the following equation [12]:

\[
\sin \theta = \frac{\lambda}{\lambda_c}
\]  

(2)

In the undertaken experiments, the waveguide simulator is assumed to be terminated with two identical unit cells, as depicted in Fig. 7. The aperture of the waveguide port is set at \( a = 30 \text{ mm} \) and \( b = 15 \text{ mm} \) to obtain a small incidence angle of 25.8° at 11.5 GHz.

The validation process is carried out in two steps. First step is the simulation of the waveguide including two identical phasing elements to obtain the phasing characteristic results in this waveguide environment. Second, the waveguide and unit cells are fabricated for measurements, as shown in Fig. 8.

Measurements are carried out on six fabricated sets of antenna elements to obtain six discrete points of phase response. These element sets have different arc angles \( \theta_s \) of 30°, 60°, 90°, 120°, 150°, and 180°. Both the simulated and measured phase characteristic results are given in Fig. 9.

From the results shown in Fig. 9, it is apparent that the simulated phase characteristic results agree well with the measured phasing result points, although small discrepancies occur when \( \theta_s \) increases from 80° to 120°.

This agreement provides high confidence in CST which is used here to generate reflection phase characteristics of the chosen phasing element.

III. FEED HORN DESIGN

In order to illuminate the reflectarray, a conical horn feed is designed and manufactured [6]. The cross-sectional configuration of the proposed horn is illustrated in Fig. 10(a). Manual optimization of horn dimensions is carried out to obtain almost a constant gain and radiation pattern over the 10.5–12 GHz band. The optimized parameters used in the manufacturing process [Fig. 10(b)] are given in Table II.

The \( E \) and \( H \) plane radiation patterns of the horn at several frequencies within the X-band are shown in Fig. 11(a) and (b). It can be seen that the measured \( E \)-plane main beamwidth is approximately 26° and remains constant for the chosen frequencies. The first side lobe of approximately \(-7 \text{ dB}\) is at ±50°. The
Fig. 11. Measured radiation pattern results of the feed at (a) E plane and (b) H plane.

H-plane main beamwidth is around 36° for all the tested frequencies. The side lobes are well below −15 dB with respect to the peak of the main beam. Therefore, stable radiation patterns in both planes are achieved by the proposed feed across the 10.5–12 GHz band. The presented results show that the designed horn is able to support wideband operation.

IV. FULL ARRAY VALIDATION

In order to benchmark the usefulness of the proposed phasing element, an offset feed reflectarray formed by 169 elements (13 × 13) is designed using the phase characteristics obtained in the previous simulations.

The unit cell spacing is chosen at 15 mm (0.58 wavelengths at 11.5 GHz). The earlier described conical horn is selected as the feed. This conical horn is assumed to be placed with 20° tilt angle in broadside direction to reduce the blockage. In order to minimize side lobes and to maximize the gain of the array, the focal length F is chosen to be equal to the reflectarray aperture size \( D = 195 \) mm. The required compensation phase value at each phasing element is computed by using the following formula from [2]:

\[
\phi_{m,n} = k_0 \left[ d_i - \left( x_i \cos \varphi_b + y_i \sin \varphi_b \right) \right] \times \sin \theta_b, \tag{3}
\]

In this equation, \( x_i \) and \( y_i \) are the coordinates of each element in \( X \) and \( Y \) directions, and \( d_i \) is the distance of each element to the phase center of the feed. The main beam of the array is assumed at \( \theta_b = 20^\circ, \varphi_b = 0^\circ \) to form specular reflection with the feed, and thus to minimize the beam squint effect. The required physical length of the variable size arc \( \{ \theta_s \} \) on each element is obtained from the characteristics at 11.5 GHz shown in Fig. 4.

It is worth mentioning that the array, which is located at 195 mm from the feeder, is located in the presented design at the border of the far-field region (207 mm) of the horn feeder. Thus, there is no need in the design and simulations to consider the technique of far-field to near-field transformation [15]. The use of that technique is effective when the array is located well within the near-field of the feeder.

The entire reflectarray is simulated in CST Microwave studio first, and then manufactured (Fig. 12) for testing. In order to minimize the computational load, the feed horn’s radiation pattern results are saved as a far-field source. The whole reflectarray antenna is assumed to be illuminated by this far-field source in CST Microwave Studio’s I-solver. Using this approach only planar reflector surface is meshed in simulations. Compared to the volume mesh of Transient Solver, the surface mesh in Integral Solver greatly reduces the number of mesh cells. The full wave simulation took one hour on a computer server of 12 cores and 48 G RAM.

Fig. 13 presents the simulated radiation pattern of the proposed reflectarray in the \( E \) plane across the frequency band from 10.5 to 12.5 GHz, as obtained from CST Microwave Studio’s I-solver. In order to minimize the computational load, the feed horn’s radiation pattern results are saved as a far-field source. The whole reflectarray antenna is assumed to be illuminated by this far-field source in CST Microwave Studio’s I-solver. Using this approach only planar reflector surface is meshed in simulations. Compared to the volume mesh of Transient Solver, the surface mesh in Integral Solver greatly reduces the number of mesh cells. The full wave simulation took one hour on a computer server of 12 cores and 48 G RAM.

Fig. 13 presents the simulated radiation pattern of the proposed reflectarray in the \( E \) plane across the frequency band from 10.5 to 12.5 GHz, as obtained from CST Microwave Studio’s I-solver. From Fig. 13, it can be concluded that the designed reflectarray points its main beam at the desired direction of 20° in the \( E \) plane with 3-dB main beamwidth staying at 9° and that the radiation pattern of the proposed reflectarray remains approximately constant across the investigated frequency band. The measured radiation patterns results of the array in \( E \) plane from 10.5 to 12.5 GHz are shown in Fig. 14. The measured results confirm that the main beam of the fabricated array precisely points at the desired 20° direction across the considered frequency band. In addition, the measured pattern of the array
stays constant in the 10.5 to 12.5 GHz band, with a 3-dB main beamwidth being approximately 9°.

By comparing the simulated and measured radiation patterns in Figs. 13 and 14, discrepancies are observed outside the main beam region. This can be explained by the fact that in simulations only a forward beam pattern of the feed is used in CST Microwave Studio’s I-solver to generate the reflectarray radiation pattern. As a result, the feeding horn’s backward and side radiations are neglected. In turn, that factor is responsible in actual radiation pattern measurements for the increased level of side and back lobes outside the main beam region. These adverse characteristics are usually part of small size reflectarray’s radiation pattern, which is the case of the present design. These undesirable effects are minimized in large size reflectarrays.

The simulated and measured gain performance is presented in Fig. 15. The proposed reflectarray has a maximum gain of 24.9 dB at 11.5 GHz. With gain drop below 1 dB, the simulated gain curve stays approximately constant between 10.9 and 13 GHz. The obtained measured gain results show that the peak value of 24.1 dB occurs at 11.6 GHz. The reflectarray’s 3-dB gain drop band is from 10.95 to 13 GHz, which is equivalent to 17.8% with respect to the central frequency of 11.5 GHz. Furthermore, the gain maintains the 3 dB gain drop bandwidth above 13 GHz, but this frequency range is beyond the operating band of the waveguide termination, which is attached to the circular feed horn. The measured gain is smaller by approximately 1 dB than the simulated one across the range from 10.95 to 13 GHz. This is due to the reasons explained with respect to the obtained simulation and experimental radiation patterns. Furthermore, the reduction of measured gain can also be due to the coaxial to rectangular and then circular waveguide transition attached to the horn.

To fully characterize the developed array, the total aperture efficiency is calculated using the method presented in [1]. From the measured performance of the array, the efficiency is found to be between 50% and 52% across the covered band. This range of values agrees well with the recent developed reflectarrays [16].

V. CONCLUSION

The design of a single layer microstrip reflectarray employing a novel phasing element to achieve wideband operation has been presented. The proposed phasing element is formed by a fixed-size circular ring with a narrow width, which is accompanied by a wide variable-length circular arc representing an open-circuited stub. To achieve a slow phase slope, thick foam is used to support the phasing element, which is developed on a thin substrate. The phasing characteristics of the proposed phasing element have been analyzed using CST Microwave Studio. The proposed phasing element offers a sufficient phase range exceeding the required 360°. An investigation into the parameters of the ring and the arc has been carried out to obtain a linear phase response when this element is placed in a unit cell. This has been accomplished by suitably selecting radius and width of the ring and width of the arc. A waveguide simulator has been manufactured to confirm the validity of the CST generated phase characteristics. The usefulness of the new phasing element has been demonstrated in the design example of an offset fed 13 × 13 elements reflectarray operating at X-band. The wideband performance of the array has been demonstrated by full wave simulations performed in the CST Microwave Studio and experimental testing carried out on the fabricated prototype. The measured results have demonstrated a 17.8% 3 dB gain-drop bandwidth confirming the wideband operation of the proposed reflectarray.

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Electronically Controlled Phasing Element for Single-Layer Reconfigurable Reflectarray

Yuezhou Li, Student Member, IEEE, and Amin Abbosh, Senior Member, IEEE

Abstract—An electronically controlled phasing element for a beam-steered single-layer microstrip reflectarray operating at 4 GHz is presented. The phasing element is formed by a printed circular ring equipped with a variable-length arc stub. The reconfigurable design is accomplished by including an open gate p-i-n transistor in the variable-length stub to offer a 1-bit beam-steering ability. Full-wave electromagnetic simulations and measurements are performed to prove the linear phase performance of the proposed cell including the effect of the utilized active chip and its biasing circuit. Those results indicate the high isolation between the phase performance of the utilized cell between the ON and OFF states of the p-i-n switches.

Index Terms—Microstrip antenna, phase shifter, reconfigurable antenna, reflectarray.

I. INTRODUCTION

MICROSTRIP reflectarray antenna is a promising substitution to the traditional parabolic reflector in terms of light weight and planar structure. It transforms the spherical wave from its feed into a planar wave in the desired direction with the help of variable-size microstrip phasing elements [1]. The proper utilization of pattern synthesis theory and careful arrangement of its microstrip elements enable the shaped beam ability of the reflectarray antenna. In particular, the use of a large number of microstrip patch elements in reflectarray offers a considerable freedom in beam shaping with much lower cost compared to the phased array. However, passive microstrip reflectarrays have the significant disadvantage of lacking the beam-steering capability. In recent literature, many ideas have been proposed to overcome this shortcoming by designing a reconfigurable reflectarray [2]–[5]. The reconfigurable design aims at combining the advantages of the fixed-beam reflectarray concerning the compact size, light weight, and low cost, with the advantages of electronic scanning in regard to increased speed compared to the mechanical steering of parabolic reflectors.

In order to electronically control the phase of microstrip elements of the reflectarray, active devices have to be included in the microstrip structures of each unit cell. Several devices are used, such as p-i-n diodes [2], microelectromechanical systems (MEMS) [3], and liquid crystal materials [4]. All these approaches are good candidates for reconfigurable reflectarray design with moderate-size apertures. However, manufacturing limitations occur for large-size arrays formed by these methods, as the control circuitry becomes highly complicated. To overcome this challenge, a multilayer aperture-coupled patch phasing element is attached with p-i-n diodes forming a reconfigurable reflectarray by gathered element method [5]. This method significantly reduces the number of the required electronic devices and reduces the complexity of the control circuitry, especially for large reflectarrays that include thousands of elements. However, the utilized multilayer aperture-coupled structure imposes other manufacturing challenges concerning the combining and alignment of the different layers. Therefore, further research work is needed to simplify the design. One obvious, but challenging, option is to use electronically controlled phasing elements in a single-substrate reflectarray.

In this letter, an electronically switchable phasing element for a single-layer microstrip reflectarray is proposed. This phasing element is formed by a circular ring with an attached variable length arc as stated in [6] and [7]. Extended investigations are carried out to improve that element’s reconfigurable phase switching capability when a practical p-i-n switch is accommodated in the variable length arc. A waveguide simulator is developed to measure the phase behavior of the reconfigurable unit cell when the attached p-i-n diode changes its state between ON and OFF.

II. UNIT CELL CONFIGURATION

The investigation is carried out on a single-layer reconfigurable reflectarray designed at C-band with a center frequency of 4 GHz. As shown in Fig. 1(a), the phasing element is designed for a wideband operation [7]. It is formed from a printed circular ring of radius $R_1$ equipped with a variable length arc stub that has a radius $R_2$. The square lattice periodicity of the array is set to $D_x = D_y = 40$ mm, which is equivalent to 0.53 free-space wavelength ($\lambda$) at 4 GHz. In the reconfigurable structure, a p-i-n switch is connected in a certain location in the arc stub across a small gap $\Delta g$ as depicted in Fig. 1(b). The proposed microstrip element is assumed to be supported by a thin substrate of thickness 0.254 mm and dielectric constant 1.96. A foam layer of thickness 3.175 mm and dielectric constant 1.06 is placed under the substrate to obtain a linear reflection phase curve as a function of the variable length arc, as shown in Fig. 1(c).

It is to be noted that $\delta_0$ and $\delta_1$ depicted in Fig. 1 are angles and not electrical lengths. The physical length of the stub can be calculated from the multiplication of its radius by the angle in radians.

In the tuning stage, the ring radius $R_1$ is fixed at 10.5 mm, and the gap ($H$) between the ring and circular arc is optimized.
The reconfigurable phasing element design is illustrated in Fig. 1(b). To obtain the beam-steering capability, the circular arc of the phasing element is divided into two parts. Part 1 is attached to the circular ring, and Part 2 forms an isolated circular arc transmission line. A p-i-n transistor switch is added across the slot between the two circular arc parts to enable the phasing elements to offer two different phase compensation values depending on the state of the p-i-n transistor. The gap of the two parts is optimized at \( \Delta \theta = 5.7^\circ \) to accommodate the utilized p-i-n transistor switch with a minimum mutual coupling between the two parts of the arc. The lengths of Parts 1 and 2 of each cell in a reflectarray are chosen in the final reflectarray design depending on the required two patterns of the reflectarray.

In order to imitate the practical case at the ON state of the p-i-n transistor, investigations are carried out by assuming three cases: microstrip line connection with no p-i-n transistor or their biasing circuit (fixed-beam case), adding the p-i-n transistor as an ideal switch with the effect of the biasing circuit, and the realistic transistor with its parasitic elements included. For the last test, the parasitic elements of Mini-Circuits PIN transistor (M3SW-2-50DR+) are extracted using Agilent ADS from the measured scattering parameters of the transistor. The obtained values are the following: resistor \( R_s = 3.1 \, \Omega \), inductor \( L_s = 0.27 \, \text{nH} \), and packaging capacitor \( C_p = 0.3 \, \text{pF} \).

The simulated phasing characteristic results of the reconfigurable element of Fig. 1(b) at 4 GHz are depicted in Fig. 3. It is clear that the inclusion of the biasing circuit and the parasitic elements of the utilized diodes have little impact on the linearity and range of the phase performance.

In another investigation, the phasing characteristic when the switch has a variable location is verified. The lengths of Parts 1 and 2 indicated in Fig. 1(b) are changed, and the phase is calculated when the p-i-n diode is in the OFF state. This case is important as it is used in the final reconfigurable design. The length of Part-1 stub [Fig. 1(b)] is increased with the total stub length

at \( 0.2 \times R1 \). The required phase linearity is obtained by making the characteristic impedance of the circular stub equal to the input impedance of the ring structure at the joint point [7]. That target is achieved by optimizing the width of the ring \( \Delta R \) and the arc \( \Delta H \). It is found that the optimum values that help in realizing a linear phasing characteristics are \( \Delta H = 6 \, \text{mm} \), and \( \Delta R = 4 \, \text{mm} \).

The reflection phase results of the phasing element are obtained by full-wave simulations of a unit cell. The frequency solver of CST Microwave Studio with appropriate unit-cell boundary conditions is applied. Here, the used assumption is that a plane electromagnetic wave is normally incident on an infinite periodic array of identical elements.

The phase characteristic results of the unit cell at the central (4 GHz), upper (4.1 GHz) and lower (3.9 GHz) frequencies of the investigated band are illustrated in Fig. 2. A sufficient phasing range of around 450° is obtained in the simulation, although with some small variations in phase when the frequency changes. The linearity of the phasing curve is validated by comparison to a linear regression curve in Fig. 2. The sufficient phasing range and the linear phasing behavior offer a lot of freedom to properly design the reconfigurable reflectarray.
remains constant such that $\theta_p = 230^\circ$, which is chosen for a maximum phase change as indicated in Figs. 2 and 3. The phase performance of this scenario is compared in Figs. 4 with the original phasing curve for the passive design (no p-i-n diodes), and with the phasing results with the switch at a fixed location and in the ON state (from Fig. 3). It is apparent from Fig. 4 that the three curves are in good agreement, although some tiny discrepancies occur when the angle is located within the 100°–180° range, which is caused by the parasitic elements of the p-i-n transistors and their biasing circuits. Therefore, the inclusion of the p-i-n switches hardly causes phase distortions, and thus they should not be observed in the operation of the reconfigurable array. Also, Part 2 of the arc has no effect on the phasing characteristics when the p-i-n transistor is in the OFF state, revealing the most required isolation between the two parts of the arc.

IV. EXPERIMENTAL VALIDATION

A waveguide simulator (WGS) [8] is used to validate the reflection phase characteristic of the electronically switchable cells. The waveguide simulator is assumed to be terminated with two identical unit cells, as depicted in Fig. 5. Because all the walls of the waveguide are conducting E-walls, the transverse-electromagnetic boundary conditions required for the correct emulation of the cell when used in a realistic reflectarray cannot be satisfied. To solve this problem, the waveguide simulator is designed to have a small incident angle. Inside the waveguide simulator, the incident angle ($\theta$) of the inbound wave on the phasing element is determined by the operating wavelength ($\lambda$) and the cutoff wavelength ($\lambda_c$) of the waveguide, as given by the following equation [8]:

$$\sin \theta = \frac{\lambda}{\lambda_c}$$

Fig. 6. Fabricated phase elements. (a) Passive phase element in waveguide simulator. (b) Full set of passive elements. (c) Reconfigurable element in waveguide simulator. (d) Full set of reconfigurable elements.

In the undertaken experiments, the aperture of the waveguide port is set at the dimensions of 80 x 40 mm$^2$ to obtain a small incidence angle of 28°.

The validation process is carried out by two steps. First, the waveguide including the two identical phasing elements as termination is simulated to obtain the phasing behavior of the structure in this waveguide environment. Second, the waveguide simulator and the unit cells are fabricated for measurement by a vector network analyzer.

Two series of phasing elements are fabricated to test the proposed cells when used in a fixed-beam and reconfigurable reflectarray. First, six sets of phasing structures are fabricated on Rogers 5880 to fit the waveguide port dimensions, as shown in Fig. 6(a) and (b). This series is measured to confirm the phasing behavior of the structure without the effect of the p-i-n transistor and biasing network at six discrete points of phasing characteristic results with the angle $\theta_p$ fixed at 40°, 70°, 100°, 130°, 160°, and 190°. The simulated and measured phase characteristic results of the passive phasing structure are given in Fig. 7. The measured phase points coincide well with the simulated phase characteristic curve and indicate a linear phase variation with more than 360° phase range.

The reconfigurable phasing elements series is fabricated on Rogers 5880LZ with Mini-Circuits PIN open gate transistor (M3SW-2-50DR+) soldered on a biasing circuit located on backside, as depicted in Fig. 6(c) and (d). This series is measured to obtain seven points of phase reflection results in both switching ON and OFF state when $\theta_p$ is chosen at 50°, 80°, 110°, 140°, 170°, 200°, and 230°.

Fig. 8 illustrates the simulated and measured phase characteristics of the reconfigurable phasing elements when the p-i-n transistor is in states ON and OFF. When the biasing circuitry is switched ON, a good agreement is obtained between the simulated and measured phase performance. A small difference can be observed between the simulated and measured performance when $\theta_p$ increases from 80° to 140°.
The reconfigurable phasing elements with p-i-n transistor in state OFF is also validated to confirm that the isolated stub Part 2 offers zero phase compensation to the steerable unit cells across a reasonable range of stub length. A good agreement between the simulated and measured results is also observed in Fig. 8. The phase of the reflection coefficient remains constant at 0° when the total central angle \( \phi_t \) is increased up to 140°. Since the length of Part 1 in the tested cells is fixed such that it has a central angle of around 50°, the achieved results mean that the maximum length of Part 2 with perfect isolation from Part 1 when the p-i-n switch is OFF is determined by 90° central angle. Using the radii and widths of the ring and stub, it is possible to verify that the electrical length of Part 2 with 90° central angle is about half-wavelength. Due to the half-wavelength resonance of the isolated arc Part 2, the sharp change in phase is observed when \( \phi_t \) is larger than 140°. This result provides a high confidence that the isolated stub Part 2 will hardly affect the performance if the length of Part 2 is controlled below half-wavelength. A length with 90° central angle means that it enables 280° phase range variation for the reconfigurable pattern as can be concluded from Fig. 4. This value is 80° less than the required full-cycle (360°) phase range. However, the 80° phase range shortage can be easily compensated when 360° phase truncation is applied to those elements that require a phase value smaller than 80°. The 90° excessive phase from the total phase range of 450° in Fig. 4 is able to accommodate the required 80° additional phase. Therefore, in the design of the reconfigurable reflectarray, phase truncation mechanism is applied to determine the lengths of stub Part 1 and Part 2 such that the length of stub Part 2 is smaller than half a wavelength.

V. CONCLUSION

A novel single-layer phasing element formed by a circular ring accompanied by a variable-length circular arc has been described for use in a reconfigurable reflectarray at C-band with 4 GHz central frequency. The beam steering is accomplished by dividing the variable-length arc into two parts and adding p-i-n switch across the slot. The performance of this reconfigurable phasing element has been validated by simulations and measurements in a waveguide simulator. The simulated phase characteristics for the reconfigurable unit cell agree well with measured values. The presented results indicate that the proposed unit-cell structure is a good candidate for a single-layer reconfigurable reflectarray design.

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Planar Array of Corrugated Tapered Slot Antennas for Ultrawideband Biomedical Microwave Imaging System

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ABSTRACT: A planar antenna array that includes 12 corrugated tapered slot elements for use in ultrawideband (UWB) biomedical microwave imaging systems is presented. The used corrugate tapered slot antenna has a compact size, low profile, moderate gain, and distortion-less performance in the time domain. The array is immersed in a carefully designed matching liquid of suitable dielectric constant to improve the matching between the array and the imaged object, and thus, to increase the dynamic range of the imaging system. A suitable platform is designed and fabricated to accommodate the array, breast phantom, and a coupling liquid for the case of UWB breast imaging. The design of the whole system is optimized using trust-region framework method in the simulation tool CST Microwave Studio. The performance of the designed array is confirmed via measurements in a realistic imaging environment.

Keywords: tapered slot antenna; antenna array; microwave imaging; ultrawideband antenna

I. INTRODUCTION

Microwave imaging modality is a promising diagnostic tool for assessing the biomedical state of the human body. One prominent example where active microwave imaging shows a great promise of such an application is early breast cancer detection [1]. The underlying notion of this technology is the significant difference of complex permittivity between malignant tumors and normal breast tissue [2].

Breast cancer persists to be the top threat to many women’s health, and early diagnosis is the key to its defeat. X-ray mammography remains the most effective screening technology for detecting clinically the breast cancer. However, it suffers from high false-negative detection rates [3]. Also, patients are exposed to ionizing radiation. Microwave detection of breast cancer is nonionizing and thus avoids the problem met in X-ray mammography [4]. In the microwave-based imaging systems, which are used in a similar fashion to the ground penetrating radar, microwave signals are transmitted from an antenna or an antenna array, and the received signals, which contain the reflections from tumors are recorded and analyzed using suitable signal processing technique to get three-dimensional images.

Many research groups have developed different types of imaging systems in recent years, such as, nonlinear inversion tomography, chirp-pulse microwave computed tomography, indirect holography, reflection-transmission holography, confocal imaging, time reversal, raster scanning, shape-based inversion, and strain imaging [5–7]. In those systems and others, different kinds of antenna arrays, such as, planar, cylindrical, hemispherical, and conformal array structures, were used to study the feasibility of detecting and localization tumor with cylindrical or hemispherical breast models [8–12]. Concerning the antenna elements used in those arrays, different types of antennas, such as, patch antenna, monopole antenna, slot antenna, bowtie antenna, and so forth, are used [13–21]. Each of these antennas has its own merits. In addition to the antenna array, a proper coupling liquid between the array and the object is necessary to reduce the reflection and enhance the dynamic range [22].
This article reports the design of a planar antenna array to be used in the microwave imaging system shown in Figure 1. In that proposed imaging system, one of the array elements is used to transmit a very narrow pulse to penetrate into the breast while the scattered signal due to different layers of breast tissues is collected by the rest of the antennas in the array. This process is repeated until all elements of the array perform the transmitting role. In this work, a planar array installed in a suitable platform is used to detect and localize breast tumor. In this technique, the breast is assumed to be inserted and slightly pressed inside the platform for a better signal penetration, and thus, more accurate postprocessing of the scattered signals.

The main challenges facing the design of suitable ultrawideband (UWB) antenna for imaging systems are the requirements for a compact size, moderate to high gain with high radiation efficiency, distortionless performance in the time domain, and low profile. To respond to those challenges, the array is built using corrugated tapered slot antennas (TSAs). This type of antenna is chosen because of its directive properties and high radiation efficiency. Because of the space limitation in breast imaging environment, a corrugation technique is used to make the antenna compact. The performance of the antenna as a single element and as an element in a planar array is tested in the presence of a breast phantom. To improve the matching between the array and breast, the array is immersed in a matching liquid that has a high dielectric constant. The simulated performance of the proposed antenna and antenna array is confirmed via measurements.

II. ANTENNA DESIGN

The antenna array is a key component in the success of microwave-based imaging systems. To achieve the best possible matching with breast tissues, the array is to be immersed in a liquid with a high dielectric constant to reduce the reflected/scattered signals at the interface and to increase the dynamic range of the imaging system as indicated in Figure 1. To efficiently use the available microwave power, directive antennas are preferred. In the proposed planar array, the elements of the array are TSAs, which have low profile and are light in weight. The requirement for one of the principles of the design of this antenna is that the end slot should reach at least one-half of the wavelengths at the lowest frequency of the desired operation. The antenna exhibits high directivity due to its traveling wave nature. It is also capable of producing a symmetric beam in the electric field plane, and the magnetic field plane, which is perpendicular to the substrate, when appropriate dimensions and slot shapes are chosen [20].

The antenna is depicted in Figure 2. It has an antipodal structure with a limited length ground plane at the bottom layer. The antenna is fed using a microstrip line of 50 Ω characteristic impedance. Initially, the proposed antenna is designed without corrugations following the guidelines described in Ref. [21]. Next, corrugations are introduced in both the top radiator and the ground layer to reduce the antenna size and to suppress the standing waves arising in the radiator of the antenna [19]. The used corrugation also

![Figure 1](image1.png)  
**Figure 1** Configuration of the microwave imaging system for breast cancer detection. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com.]

![Figure 2](image2.png)  
**Figure 2** Configuration of the designed antenna: (a) without covers and (b) with covers. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com.]
improves the directivity in the lower part of the UWB frequency range and enhances transmission of UWB pulses. The top radiator and the bottom ground are covered by a dielectric material with properties that are similar to those of the substrate to protect the radiating element from the adverse effects of the coupling liquid, such as metal corrosion, and to improve the matching of the antenna when immersed in a coupling liquid.

The lowest frequency of operation (\(f_1\)), substrate thickness (\(h\)), and the dielectric constant (\(\varepsilon_r\)) is used to calculate the width (\(w\)) and the length (\(l\)) of the antenna’s structure in free space

\[
w = l = \frac{c}{\sqrt{\varepsilon_r} f_1},
\]

where \(c\) is the speed of the light in free space.

The intersection of quarter of two ellipses is used to form the structure of the radiating elements. The width of the microstrip transmission line \(w_f\) is calculated using Eq. (5) to get characteristic impedance \(Z_0 = 50 \Omega\).

\[
z_0 = 60\sqrt{(\varepsilon_r) \ln[(8h/w_f + w_f/4h)]}
\]

Assuming that the antenna and array are to be designed for the frequency range from 3.1 to 10.6 GHz according to the FCC regulations and the IEEE recommendations for medical imaging systems, the lowest frequency of operation (\(f_1\)) is then chosen to be 3.1 GHz. The length of the slots used in the corrugations, shown in Figure 2, is chosen to be quarter wavelength at the center frequency of operation (6.85 GHz).

The final dimensions are obtained using the optimization capability of the software CST Microwave Studio. Trust-region framework method is used to optimize the design parameters. The main feature of that method is that it uses the sensitivity information of the different design parameters to cut down the optimization time dramatically. It is used in this work to watch the return loss of the antenna and to find the optimum dimensions for the antenna and slots with 10-dB return loss as a reference across the band from 3.1 to 10.6 GHz. The optimized dimensions of the antenna assuming the use of Rogers RT6010 with dielectric constant = 10.2 and thickness = 0.635 mm as the substrate are shown in Figure 2 are \(L = 40\), \(w = 22\), \(L_s = 4\), \(W_s = 1\), and \(S_p = 0.5\) mm.

III. MATCHING MEDIUM DESIGN

The antennas in the imaging system, depicted in Figure 1, are immersed in a liquid with a high dielectric constant to achieve the best possible matching with the human tissues [22]. Also, the use of high dielectric constant matching liquid enables reducing the size of the antennas. The dielectric properties of different mixtures of materials that are selected as possible matching liquid are tested using the HP85070B coaxial probe connected to the HP network analyzer (HP8530A). The probe provides the information about the real and imaginary parts of the complex permitivity of a tested material across the frequency range from low microwave frequencies (1 GHz) to about 18 GHz. The complex dielectric constant is represented by the following expressions:

\[
\mathbb{C}(\omega) = \mathbb{C}'(\omega) + j\mathbb{C}''(\omega)
\]
\[ \varepsilon(\omega) = \varepsilon_{\infty} + \left( \frac{1}{\Delta\varepsilon} + j\omega \tau \right) + \frac{j\sigma_s}{\omega\varepsilon_0} \]  \hspace{1cm} (4)

The conductivity can be related to the imaginary part of the complex dielectric permittivity as follows:

\[ \sigma_s(\omega) = \omega\varepsilon_0 \varepsilon''(\omega) = 2\pi\sigma_0 \varepsilon'' \] \hspace{1cm} (5)

where \( \omega \) is the angular frequency, \( \varepsilon'(\omega) \) is the frequency-dependent dielectric constant (real part of complex permittivity), \( \varepsilon_0, \varepsilon''', \Delta\varepsilon, \sigma_s, \tau \), and \( x \) are the fitting parameters used in the HP85070B coaxial probe.

The full-wave analysis performed in Ref. [22] using a multilayer model of breast with a normal wave incidence provides a guideline on the required properties of a matching medium that can reduce the adverse effects of signal reflections at the antenna–breast interface. Those guidelines are used to design a suitable coupling medium between the antenna and the breast. A mixture of glycerine, water, and corn syrup is chosen to meet these objectives. Figure 3 shows the measured variation of the dielectric permittivity with frequency for the proposed matching liquid which is a mixture of 1:0.5:2 of glycerine, water, and corn syrup.

To refine the design of the antennas in the presence of the manufactured matching liquid, the obtained measured properties of the developed liquid are used as an input data to the simulation tool CST Microwave Studio during the design and analysis of UWB antenna elements and arrays operating in the presence of a coupling medium. The effect of dielectric scaling on size, input matching, and radiation pattern is evaluated using the software.

IV. ANTENNA’S PERFORMANCE IN MATCHING LIQUID

Performance of the proposed antenna is first verified via computer simulations. Next, the antenna is manufactured (Fig. 4) and tested to confirm its simulated performance. The simulated and measured results with and without the coupling liquid are shown in Figure 5. The obtained results indicate that when the coupling medium is used, the antenna features UWB performance from 3.1 to 10.6 GHz assuming the 10-dB return loss as a reference to define the bandwidth specification. It is to be noted that the difference between the simulated and measured results in Figure 4 is due to the effect of the multiple reflections at the boundaries of the plastic container used for the coupling liquid, and that container is not included in the simulations whereas its effect is obviously included in the measured results.

To confirm the effect of using the designed coupling liquid on the performance of the antenna, the return loss is calculated in free space, that is, without coupling liquid. It is clear from the results depicted in Figure 5 that the performance deteriorates dramatically especially at the lower part of the band.

To verify the directive properties of the designed antenna, the maximum gain is calculated. It is found to be from 2.8 to 5.7 dBi across the band from 3.1 to 10.6 GHz. In comparison with the gain of other designed

![Figure 5](image-url) Performance of the antenna with and without coupling liquid. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com.]

![Figure 6](image-url) The time domain response of the antenna: (a) simulated and (b) measured. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com.]

![Figure 7](image-url) The calculated group delay of the antenna. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com.]
antennas [13–21], this is an acceptable medium range of gain values considering the very compact size of the antenna.

To verify the capability of this antenna to support transmission and reception of narrow pulses in a distortionless manner, the time domain impulse response of the antenna is measured. In this case, two antennas are put at the same height above the ground and distance of 30 mm between them. A very narrow pulse, which has UWB frequency contents, is generated using the vector network analyzer (R&S ZVA24). The pulse is then received using the other antenna. The result of measurement, when the two antennas are aligned face to face is shown in Figure 6. It is clear that the developed antenna supports almost distortionless transmission which is very important in a microwave imaging system so that ghost targets are avoided. To confirm the distortionless performance of the antenna in the presence of the designed coupling liquid, the group delay between two similar antennas at a distance of 30 mm is calculated. The results depicted in Figure 7 reveal a very low distortion across the band of interest.

V. ARRAY AND PLATFORM DESIGN

A planar array comprising of $6 \times 2$ corrugated TSAs is built in the manner shown in Figure 7. The antenna elements are supported by a plastic material with relative dielectric constant equal to $\varepsilon_r = 3.1$.

The two important parameters that define the effectiveness of the used array are the return loss of each of the elements that form the array and the level of mutual coupling between the different elements. The horizontal space between the element ($h_0$) and vertical space ($v_0$), which are shown in Figure 8, are optimized for minimum values that make the mutual coupling between the array elements less than $-20$ dB and the return loss of each element more than 10 dB assuming the array is immersed in a coupling liquid and in front of the imaged breast.

Trust-region framework method in the software CST Microwave Studio is also used here to watch the mutual coupling between each pair of the array elements and to find the minimum distance between the elements of the array with mutual coupling less than $-20$ dB as a
The worst case of the coupling liquid occurs between the nearest elements in the array. The optimized values for \((h_s)\) and \((v_s)\) are found to be 19 and 4 mm, respectively. The platform that supports the planar array is designed using a box with dimensions of 250 × 300 × 150 mm\(^3\) width, length, and depth, respectively. It has a stand to hold the breast phantom that has a semirectangular shape due to a slight compression by the platform’s lower plate. This stand can be moved up and down to scan the whole phantom and also it is adjustable so that the distance between the array and the phantom can be controlled for the optimum value. The configuration of this system is shown in Figure 9.

VI. ARRAY’S PERFORMANCE

To confirm the performance of the array, a prototype is manufactured and tested. The test is performed when the platform is filled with the designed matching liquid. The worst case of mutual coupling between the different elements of the array is shown in Figure 10. It is clear that the array has a good performance concerning the level of mutual coupling between any pair of antennas which is less than −25 dB across the band from 3.1 to 10.6 GHz.

To test the array in a realistic environment, a heterogeneous breast phantom that has average electrical properties equal to those of real breast tissues is designed and manufactured. It has the dimensions of 110 × 10 × 10 cm\(^3\) width, height, and depth, respectively. To manufacture the phantom, 200 ml grape-seed oil, 25 g propylene glycol, 193 mL milli-Q water, 40 g gelatine, 1.512 g of 32% formalin solution, and 2.2 mL commercial dishwashing liquid as a surfactant to form oil emulsion are used [23]. The quantity of the propylene glycol and gelatine is changed during the manufacturing process to have a phantom permittivity and conductivity close to the realistic values of a low dense breast that includes mostly fatty tissues [1, 24]. To make the developed phantom heterogeneous, the propylene glycol and gelatine are distributed randomly. Figure 11 shows the measured variation of the dielectric permittivity and conductivity with frequency for the manufactured phantom. Those values are close to the values for realistic healthy breast tissues [1]. Figure 12 shows the manufactured phantom when it is placed in the platform without and with the coupling liquid. It is worth mentioning that the developed phantom has a rectangular shape because the breast is assumed to be slightly pressed for a better imaging using the designed planar array.

The phantom is inserted inside the platform as shown in Figure 12 and the antenna elements are tested again. In this test, the return loss is measured in two cases. In the first case, the coupling liquid is not used, that is, a free space is available between the array and the phantom. In the second case, the designed coupling liquid is used to fill the platform. The results depicted in Figure 13 reveal the low values of the return loss especially at the lower part of the band when no coupling liquid is used. This is due to the strong backscattering of the transmitted signal at the free space–phantom interface. Those undesired
scattered signals are significantly reduced, and the return loss is increased when the coupling liquid is used due to the perfect matching at the interfaces of antenna–coupling liquid–breast phantom.

To assess the distortion level in the transmitted pulses inside the breast model, the fidelity factor is calculated at different locations within the breast phantom. The parameters of the whole imaging system (antenna array, platform with breast phantom, and matching liquid) are included in the CST Microwave Studio to calculate the fidelity factor. This factor is defined as the maximum magnitude of the cross-correlation between the observed pulse at a certain distance and the excitation pulse [25]. The result is shown in Figure 14 where the effect of all the scattered/reflected signals is included. The results show that as the signal propagates through the human body, the fidelity factor is decreased. This indicates an increasing pulse distortion inside the breast. For the antenna presented in this article, Figure 14 shows that the fidelity factor inside the breast phantom in the end-fire (normal) and off-normal directions is within a reasonable range that guarantees clear post-processing images.

VII. CONCLUSION

The design of a directive UWB antenna array immersed in a coupling liquid for use in an UWB biomedical microwave imaging system has been presented. To minimize the size of the antenna, corrugations have been introduced in both the top radiator and the ground part of the antenna. To protect it from the adverse effects of the coupling liquid (such as corrosion of metallic parts), the radiator and the ground are covered by a dielectric material that has dielectric properties similar to those of the antenna’s substrate. The designed array has been tested with respect to a particular application of breast imaging. To this purpose, a special platform accommodating the array, matching liquid, and breast phantom has been constructed.

The measurement and simulation results have shown that the return loss of each of the antenna elements in the array covers the band from 3.1 to 10.6 GHz assuming the 10 dB return loss as a reference. Moreover, the level of the mutual coupling between any pair of antennas is less than −25 dB across the microwave band measured in the absence of the breast phantom.

The time-domain performance of the antenna shows its capability to support transmission and reception of narrow pulses without distortion.

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macroyclic complexes. A short period shared between
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Experimental investigations into detection of breast tumour using microwave system with planar array

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Abstract: A microwave imaging system for the potential detection of tumours in breast tissues using a planar array that employs 6 $\times$ 2 compact tapered slot antenna elements is reported. In this system, the breast is placed on a plastic sheet and slightly compressed to form a semi-rectangular shape; this allows for accurate image reconstruction using a planar array. Both the array and the phantom are immersed in a coupling medium to increase the signal penetration, and thus, the dynamic range of the system. In order to quantify the effect of changing the number of elements and their positions in the array, image reconstruction is undertaken with three different configurations. That is with 6- and 12-array elements and an additional 90° rotation of the phantom. To test the designed system a suitable combination of materials are used to fabricate a low-density, heterogeneous breast phantom. Two- and three-dimensional versions of a confocal imaging algorithm are used to construct the images and metrics are proposed to evaluate the quality of the image. Single and multiple embedded targets are shown to be resolved using the proposed array and imaging algorithm in either two or three dimensions.

1 Introduction

Microwave imaging is a potentially new method for breast cancer diagnosis \cite{1}. Ultra-wideband (UWB) microwave imaging systems have been investigated using different types of approaches depending on how the data are collected and processed. These systems can be monostatic, bistatic or multistatic. For the monostatic method, one antenna is used as both a transmitter and receiver and moved across the breast to form the synthetic aperture. In the case of bistatic, one antenna is used as a transmitter and another antenna as a receiver, whereas in the multistatic case, an array is used to collect the data.

Different configurations of arrays have been used in microwave imaging systems, such as circular, planar, conformal and cylindrical arrays \cite{1–7}. In these arrays, different antenna configurations have been used to form the arrays such as bowtie, dipole, microstrip, dielectric resonator, slot and horn antennas \cite{1–11}.

In this work, a microwave imaging prototype employing a planar array of corrugated tapered slot antennas that are immersed in a coupling medium is reported. In this system, each antenna has the ability to transmit an UWB signal and receive the backscattered signal. To reduce the unwanted reflections and increase the dynamic range, the array and the imaged breast are placed in a carefully fabricated coupling medium. A suitable mixture of materials is used to fabricate a heterogeneous phantom that is needed to test the designed system. In the proposed system, the breast is assumed to be gently pressed so that it can take the shape of semi-rectangular prism. This approach enables the use of a planar array and the associated simple post-processing algorithms. In this work, two-dimensional (2D) and three-dimensional (3D) confocal imaging algorithms are used.

2 Overview of the microwave imaging system

Fig. 1 shows the microwave imaging system. The system consists of a 30 $\times$ 25 $\times$ 15 cm rectangular box as a platform for imaging, electronic sub-system using two single-pole-eight-throw electro-mechanical coaxial switches, a vector network analyser (VNA) with time-domain processing capability, a low dense heterogeneous breast phantom compressed to semi-rectangular shape and a personal computer (PC) for control of measurement, data storage and processing. The platform consists of planar array supported by a plastic sheet and fixed in the platform with a lift that allows the mechanical rotating and scanning of the phantom. The platform is filled with a fluid designed to have certain electrical proprieties for optimal coupling between the antenna and the breast phantom.

Each group of six elements of the array is connected to the output port of electro-mechanical coaxial switches (model 50S-1317) offering more than 70 dB isolation and 0.2 dB insertion loss across the 3–11 GHz band. The switches operate at 12 V and require 325 mA of current. The collection of either the time or frequency domain scattering data from all elements of the array is achieved by activating the VNA from the control PC.

A low-dense heterogeneous breast phantom compressed to 10 $\times$ 11 $\times$ 10 cm semi-rectangular shape with dielectric...
permittivity that changes from $\varepsilon = 12$ to 8, and conductivity that changes from $\sigma = 0.5$ to 2.15 S/m across the band from 3.1 to 10.6 GHz is manufactured. Two hundred millilitres grape seed oil, 25 g propylene glycol, 193 ml milli-Q water, 40 g of gelatine, 1.512 ml of 32% formalin solution and 2.2 ml commercial dishwashing liquid as a surfactant to form oil emulsion are used to manufacture the phantom [12]. The quantity of the propylene glycol and gelatine is changed during the manufacturing process to have a phantom permittivity and conductivity close to the realistic values of a low dense breast that includes mostly fatty tissues [1,13]. To make the developed phantom heterogeneous, the propylene glycol and gelatine are distributed randomly. A tumour with average dielectric proprieties ($\varepsilon = 54$ and $\sigma = 2.2$ S/m) is also manufactured using 200 g gelatin, 328 ml deionised water, 17 ml $n$-propanol, 0.346 g $p$-toluic acid, 3.72 g formaldehyde, 30.4 ml safflower oil and 2 ml commercial dishwashing liquid [14]. A straw with 0.5 mm radius and 25 mm height is filled with the manufactured tumour and inserted inside the manufactured breast phantom.

In order to improve the matching between the antenna array and the breast phantom that can result in a significant improvement in the dynamic range of the imaging system, a coupling liquid is manufactured and used to fill the platform. The electrical properties of the liquid are chosen according to Abbosh et al. [15]. This mixture is formed using 1:0.5:2 of glycerine, water and corn syrup. The average dielectric permittivity and conductivity of the liquid is 10.4 and 1.98, respectively, at the centre frequency 6.85 GHz.

3 Design and performance of the array

Antennas of antipodal tapered slot structure with outer edge corrugations in both layers are used to form the array. The antipodal structure of the radiating element is elliptically tapered [16]. The corrugations in both the top radiator and background are introduced to reduce the size of the antenna and to improve its directivity in the lower part of the UWB frequency [17]. The length of the slots used in the corrugations is chosen to be a quarter-wavelength at the centre frequency of operation, that is, 6.85 GHz.

The guidelines in [16,17] are used to design the antenna, which is fabricated (Fig. 2) using Rogers RT6010 with dielectric constant = 10.2 and thickness = 0.635 mm as the substrate. Giving the properties of the utilised substrate (thickness $h$, and dielectric constant $\varepsilon_r$) and the lowest frequency of operation $f_1 = 3.1$ GHz, the initial values for the width ($W$) and length ($L$) of the antenna are calculated using the following equation [16]

$$W = L = \frac{C}{f_1 \sqrt{\varepsilon_r + \frac{2}{\varepsilon_r + 1}}}$$  \hspace{1cm} (1)$$

where $C$ is the speed of light in free-space.

The dimensions of the antenna are then optimised using the simulation tool CST Microwave Studio. The optimum dimensions are $L = 40$ mm, $W = 22$ mm and the length of corrugations = 4 mm. The top radiator of the antenna and the background metal are covered using a material that has dielectric properties similar to the utilised substrate to protect the antenna from the adverse effects of the coupling liquid and to maintain optimal matching with the coupling liquid. As indicated in Fig. 3, the antenna when immersed in the designed coupling liquid has more than 10 dB return loss across the band from 3.1 to 10.6 GHz. To verify the
The microwave imaging set-up data for this system are provided in Fig. 5. The body to be imaged is assumed to be cuboid-shaped that is discretised at evenly spaced points. The points that lie on the surface are denoted by \( E_i(x, y, z) \). Any discrete point inside the surface is denoted \( p(x, y, z) \). The body to be imaged has permittivity and conductivity that are denoted \( \epsilon_{ph} \) and \( \sigma_{ph} \), respectively. The coupled medium surrounds the body to be imaged has permittivity and conductivity of \( \epsilon_m \) and \( \sigma_m \), respectively. The 12-antenna elements of the planar array are used as signal.

**Fig. 4** 3D radiation patterns of the antenna at

\( a \) 3.1 GHz

\( b \) 10.6 GHz

4 Data acquisition

The experimental prototype system shown in Fig. 1 is used in the monostatic and bistatic modes of operation. The monostatic mode is operated when the six-antenna elements in the upper or the lower part of the planar array is used to image the object. The bistatic mode is operated when the mechanical movement and the mechanical rotation facility of the lift is used to move the phantom up and down and rotate it by 90°.

The UWB pulses are generated in both modes using VNA in a step-frequency manner across the band from 3.1 to 10.6 GHz. In the first case, the data (complex \( S \)-parameters) are collected by activating either port 1, which is connected via a switch to the upper six elements of the array or port 2, which is connected to the lower six elements of the array, of the VNA to record the data from the six-antenna elements in the upper or the lower part of the planar array \( S_{11}, S_{22}, \ldots, S_{66} \) or \( S_{77}, S_{88}, \ldots, S_{1212} \). The monostatic approach is chosen here as it has so far given satisfactory results and avoids the complexity in hardware and software required for a multisistatic approach.

The body to be imaged (rectangular breast phantom) is placed on the adjustable lift. The horizontal distance between the antenna and the phantom is fixed to 20 mm, whereas the vertical distance is adjusted to scan different slices from the phantom (the section with target, and the section without target). In the data acquisition, the data are first collected from port 1 for the upper part of the array that faces a certain upper slice of the phantom. The lift is then moved vertically to a distance equal to 26 mm (the vertical distance between the two lines of the array) to ensure that the antennas in the lower part face the same slice faced previously by the upper part of the array. The data are then collected from port 2. The data from ports 1 and 2 are combined. In this case, data from 12-antenna elements at different locations (as the lower part of the array is shifted horizontally with respect to the upper part as indicated in Fig. 1) with respect to the phantom are recorded to image the object. Finally, the bistatic mode is repeated when the phantom is rotated 90° to allow ports 1 and 2 of the VNA recorded the data of 12-antenna elements after the phantom rotated by 90°.

In the previous two modes, all the antenna elements are used as transmitters and receivers. Initial data for the two ports \( (S_{11}, \ldots, S_{1212}) \) of the VNA are recorded when there is no imaged object present in the platform. These data are subsequently used to remove reflections caused by the platform.

5 Post-processing algorithm

In this work, a delay-and-sum confocal algorithm [19] with a slight change is used to reconstruct the rectangular-shaped breast phantom. In order to create accurate images, it is necessary to find the correct path that the wave-front travels. For that reason, Fermat’s principle [20] is used to estimate the path of the wave. It states that the path that results in minimal propagation time is the real path. The electromagnetic signals travel in the coupling medium and the heterogeneous phantom; therefore the average dielectric proprieties of the imaged body must be estimated to accurately predict the length of the electrical path of the signal.
sources. Before the confocal process can be applied, it is necessary to perform the following pre-processing steps:

1. \( N \) time domain signals \( A_n(t) \) should be obtained, where \( n = 1, 2, 3, \ldots, N \), where \( N = 12 \) in this work.
2. The scattered signals can be obtained from the incident and the total field \( A_n^{\text{scatter}}(t) = A_n^\text{total}(t) - A_n^{\text{incident}}(t) \) where \( n = 1, 2, 3, \ldots, N \).
3. To cancel any background signals such as the reflections from the skin layer the differences in signals between the antennas are constructed.

For \( n = 1, 2, 3, \ldots, N - 1; F_n(t) = A_n^{\text{scatter}} - A_{n+1}^{\text{scatter}} \) and \( F_N(t) = A_N^{\text{scatter}} - A_1^{\text{scatter}} \).

In order to mathematically explain the iterative process of the imaging algorithm, it is convenient to define several mathematical sets. Referring to Fig. 5, we consider the object to be imaged to be discretised into points denoted \( p \) defined in Euclidean space by \((x, y, z)\) coordinates. Boundary points that exist on the edges of the body to be imaged are denoted as \( B_p = \{ B_1(x, y, z), B_2(x, y, z), B_3(x, y, z), \ldots, B_{N_b}(x, y, z) \} \) where \( B_i(x, y, z) \) is the \( i \)th boundary points in rectangular space defined by \( x, y \) and \( z \) coordinates, and \( N_b \) is the number of the boundary points corresponding to the skin layer. The set which contains all points in the body and on the surface is denoted \( Z \). The antenna spatial coordinates are also denoted as \( \{ A_1 = A_{a1}(x, y, z), A_{a2}(x, y, z), A_{a3}(x, y, z), \ldots, A_{a65}(x, y, z) \} \) where \( A_i(x, y, z) \) is the \( i \)th antenna and \( N \) is the number of antennas. The elements in the top array are denoted by \( A_{a1}, A_{a2}, \ldots, A_{a65} \), whereas lower array elements are \( A_{a7}, A_{a8}, \ldots, A_{a12} \). The pseudo-code for the confocal imaging algorithm is Fig. 6.

Here we implement Fermat’s principle by constructing all possible propagation paths from the antenna to the boundary points, and then from the boundary points to the current point \( p \); accordingly, our optimal path is the distance and the velocity of the wave in the coupling medium and in the phantom. A continuous colour image is produced using a shading operator to interpolate at non-tested points. Strong intensity colours indicate the location of significant scattering objects. In this work, the average value of the permittivity and the conductivity for breast phantom are set to \( \varepsilon_p = 10 \) and \( \sigma_p = 1.3 \), respectively. The background coupling medium has \( \varepsilon_m = 10 \) and \( \sigma_m = 1.3 \), respectively. The abovementioned algorithm gives the iterative process to construct the image from the processed scattered data. The colour map intensity denoted by \( I \) is given as a function of \( p \).

To evaluate the effectiveness of the produced images quantitative metrics are used [19]. In order to explain the metrics used, it is convenient to define a further set of points \( \tau \) that map to the location of the emulated tumour in the phantom. The first metric is the ratio of the average intensity value of points located in the tumour area divided by the average intensity points in normal breast tissues denoted \( Q \) and defined as

\[
Q = \frac{\mu[I(p)]}{\mu[I(p) \notin T]} \quad \forall p \in T
\]

where \( \mu[\bullet] \) is the mean function. A higher value for this metric means the intensity of the tumour region is larger than the background regions. The second metric represent the ratio of the maximum intensity value of the tumour area over the maximum intensity of the remaining points in the colour map, denoted by \( \gamma \), and given as

\[
\gamma = \frac{\max[I(p)]}{\max[I(p) \notin Z]} \quad \forall p \in T
\]

The third metric is the absolute distance between the location of the tumour and the location of the maximum intensity given in the reconstructed image. If \( t \) denotes the centre of the tumour, this metric is defined as the following

\[
E = |p^* - t| \quad \text{where} \quad p^* = \operatorname{argmax}[I(p)] \quad p \in T
\]

6 Result and discussions

The designed system is used to image the manufactured cuboid-shape breast phantom. All the manufactured tumour samples are inserted inside a 5-mm diameter straw with 25-mm length. The imaging algorithms are explained in the next section with 2D and 3D versions.

6.1 Two-dimensional image reconstruction using six-antenna elements

A 2D image reconstruction with the six-element array is done by collecting the time-domain data from port 1; subsequently the time-domain S-parameters \( S_{11}, S_{22}, \ldots, S_{66} \) are passed to the imaging algorithm. The emulated breast phantom and tumour is placed 20 mm away from the centre of the upper
part of the planar array. The phantom is centred to face the array. The target is inserted inside the breast phantom at the location 35.5, 66.5 and 50 mm; this position is denoted T1. The algorithm was set to compute the image intensity at 2618 evenly spaced \((x, y)\) points at 1-mm intervals. On average, this took 30 min to complete on a PC with Intel Xeon X5650 processor and 48 GB of RAM. Fig. 7a shows the imaging result of this configuration. It is clear that the system is successful in detecting the tumour. Another configuration is subsequently imaged with two emulated tumours at (24, 54 and 35 mm) and (80, 83 and 35 mm) denoted T2 and T3, respectively. The result is shown in Fig. 7b. It is clear from the results that the two targets are successfully detected.

In order to verify that using directive antennas in the array is necessary to resolve targets in the \(z\)-dimension, a test of the phantom with the target is done by moving the phantom up or down a distance equal to 35 mm. This distance is measured to ensure that we are scanning the area in the breast phantom above or below the target level. The process in this section is repeated and the imaging result of the breast phantom upper, and similarly lower, the target is shown in Fig. 7c. It is clear that this region appears without a scattering object.

6.2 Two-dimensional image reconstruction using 12 elements

A 2D image reconstruction of 12-antenna element array is done by collecting the data from the upper and the lower part of the planar array. The time domain scattered signals \(S_{11}, S_{22}, \ldots, S_{66}\) are captured by port 1 of the VNA, whereas the signals \(S_{77}, S_{88}, \ldots, S_{1212}\) are captured by port 2 and passed to the imaging algorithm. The imaging process is repeated for the same targets T1, T2 and T3 mentioned in the previous section. A further emulated tumour was also used for this experiment at position (66, 54 and 50 mm) denoted T4. This is to verify the ability of the planar array to detect targets that are further away from the array.

The phantom is moved down using the lift by a distance of 26 mm to ensure the two lines of the array face the same section of the phantom. The distance between the array and the phantom is kept at 20 mm.

Image reconstruction for phantoms with T1 and T4 are shown in Figs. 8a and b. Image reconstruction of phantoms with the dual targets (T2 and T3) is given in Fig. 8c.
Comparing the results of using the 12 elements of the array (Figs. 8a and c) as opposed to the six-element array (Figs. 7a and b) indicates that the background intensity of the non-tumour regions decreases when the number of elements in the array is increased.

### 6.3 Two-dimensional image reconstruction using 12 elements with 90° phantom rotation

This experiment involves using the 12-element array and recording the S-parameters at 90° phantom rotation. This is done by collecting the data from port 1 as mentioned in section one and the phantom is subsequently rotated 90° to collect the data again from the same port, but in this case the antenna array faces a different angle of the phantom. Therefore there are two sets of S-parameter data: with and without phantom rotation. The phantom is moved down as in Section 2 and the data are collected again from port 2 with and without phantom rotation. To set the imaging, we combine the data without rotation in the first and in the second group together and the data with the rotation. In this case, we have 12 datasets without phantom rotation and 12 datasets with phantom rotation. These datasets are used to compute the image of the phantom with T1, T2, T3 and T4 targets as shown in Figs. 9a–c. Compared with the previous imaging results, it is clear that the resolution of the target increases when the phantom rotation is done and the and the background intensity of the non-tumour regions again has decreased.

### 6.4 Three-dimensional imaging results

The previous results only consider image reconstruction in a 2D plane because it allows results to be obtained in a short period of time. We can, however, reconstruct the images in a 3D using the same algorithm. In this experiment, the algorithm computes the image intensity at 17.501 (x, y, z) points spaced at 4-mm intervals in the x, y and z dimension. The body to be imaged was illuminated at different z-axis levels ranging from 10 to 90 mm at 10 mm intervals. On average, it took 6 h to complete the image reconstruction process. Fig. 10 gives 3D images produced by the presented algorithm. These images were produced by plotting voxels centred at the particular x, y, z coordinate when the intensity at that point was within 90% of the maximum intensity.

The figure shows that the proposed algorithm is successful in resolving single and multi-targets in the heterogeneous breast phantom.

### 6.5 Metrics of the reconstruction images

The metric parameters are calculated for the imaging results of the explained three cases to evaluate the accuracy of the three utilised imaging techniques. Table 1 gives the calculated value of \(Q\), \(γ\) and \(E\) parameters for all the three cases. For the results shown in Fig. 7, the image intensity as depicted by \(Q\) for T1, and T2 in the specific area of the tumour is 1.8 times more than the intensity of the background. For, T3 it is 0.53. In the same figure, the \(γ\) parameter is 1 for both T1, and T2, whereas it is 0.45 for T3. These results indicate that the tumour is the strongest scattering object when it positioned close to the array, whereas its strength as a scattering object diminishes when it is positioned away. The \(E\) metric for T1 and T2 is much smaller compared with T3 which is very high. The \(E\) metric indicates that the difference in the centre position of the real target and detected target. The differences are \((T1) = 1.15\) mm, \((E(T2)) = 1\) mm and \((E(T3)) = 6.5\) mm.

For the results of Fig. 8 where the 12-antenna elements are used to generate the images, Table 1 shows that the \(Q\) factors for T1, T2 and T3 are 2, 2 and 0.77 indicating that the tumour behaves as a stronger scattering object when the number of the antenna is increased. The parameter \(γ\) is one for all T1, T2 and T4, whereas \(γ\) for T3 is equal to 0.59. These results indicate an improvement in the detection capability when compared with using only 6-antenna elements. The \(E\) metric for the results of Fig. 8 are \((T1) = 1.1\) mm, \((E(T2)) = 1\) mm and \((E(T3)) = 4.5\) mm. These values indicate an improvement in predicting the correct position of the tumours even if they are close to the centre position of the phantom when the number of utilised antennas increases.

For the imaging results of Fig. 9 that are obtained by using the combination of 12-antenna elements and mechanical rotation, the \(Q\) parameter has the same values as in the technique that does not involve mechanical rotation for the targets T1 and T2. For T3, the metric \(Q\) is improved compared with the two previous techniques. The metric \(γ\) for T1, T2 and T4, is the same as previously calculated while it is improved to 0.7 for T3. The parameter \(E\) in this technique for the different targets are \((T1) = 0.83\) mm, \((E(T2)) = 0.85\) mm, \((E(T3)) = 3.2\) mm and \((E(T4)) = 1.2\) mm. In comparison with the other two techniques that do not include mechanical rotation, it is clear that the difference in the positions of the real target and detected target is much smaller. All these results indicate that phantom rotation with larger number of antennas produces superior images.

### 7 Conclusion

The capabilities of a microwave imaging system that employs a \(6 \times 2\) planar array for breast tumour detection have been demonstrated. In the designed system, the breast is assumed to be gently pressed so that it can have a cuboid shape. Three different cases are used to collect the data: 6-antenna element, 12-antenna elements and 12-antenna elements with 90° phantom rotation. A delay-and-sum confocal algorithm

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**Table 1** Performance of the algorithm for the three reconstructing image methods

<table>
<thead>
<tr>
<th>Targets</th>
<th>6-antenna elements</th>
<th>12-antenna elements</th>
<th>12-antenna elements with phantom rotation</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>(Q)</td>
<td>(γ)</td>
<td>(E,) mm</td>
</tr>
<tr>
<td>T1</td>
<td>1.8</td>
<td>1</td>
<td>1.15</td>
</tr>
<tr>
<td>T2</td>
<td>1.8</td>
<td>1</td>
<td>1</td>
</tr>
<tr>
<td>T3</td>
<td>0.53</td>
<td>0.45</td>
<td>6.5</td>
</tr>
<tr>
<td>T4</td>
<td>1.85</td>
<td>1</td>
<td>1.25</td>
</tr>
</tbody>
</table>
is used to reconstruct 2D and 3D images of a rectangular shaped heterogeneous breast phantom. The results show the ability of the system in detecting a single or multi-tumours in different locations.

## References

The degradation of the sidelobe level is below 8 dB. The degradation of the sidelobe level is due to the perturbation caused by the WR10 waveguide sections and the transitions used in the measurement. Measurements always include the transitions and an accurate measurement of the antenna gain is difficult. The measured gain, including the losses due to the transitions, varies from 7.59 dBi at 88 GHz to 5.75 dBi at 90 GHz. Fabrication should avoid any air gap between the dielectric and the metal plate [7]. Simulations show that even a 2 mil air gap significantly affects the leaky mode excitation and, as a result, the leakage from the mode may disappear.

V. EFFECT OF DIELECTRIC AND METAL LOSSES

At millimeter wave frequencies, the dielectric and metal losses play an important role in the gain performance of an antenna by reducing the antenna efficiency. Simulated antenna gain, under different loss conditions, at 88 GHz, are given in Table IV. The same slot antenna discussed in Section IV, with copper, and the loss tangent of the substrate at 0.0001, but without the WR10 to SINRD waveguide transition, are considered. The back-to-back WR10 to SINRD transition loss is within 0.5 dB. Due to the metal loss, the gain is reduced by 0.91 dB from its original value of 13.20 dBi. It is reduced by 0.35 dB due to the dielectric loss. Finally, when all losses are considered (tip metal losses + metal losses + dielectric loss), the antenna gain is reduced by 1.19 dB and the corresponding antenna efficiency is 37%. However, these values depend upon the materials and the length of the antenna.

VI. CONCLUSION

The operating bandwidth of an SINRD waveguide depends on the periodicity of the hole arrays. As a result, the fast-wave frequency band also depends on periodicity. The bandwidth increases with decreasing periodicity. It can be utilized to tune the scan rate of an LWA. There is no need to change the substrate specifications. The LWAs are electrically long antennas (several wavelengths). But, these antennas are easily implemented in millimeter wave band due to the small wavelength at this frequency range. In this context, it is necessary to implement the antenna using a low-loss dielectric material.

ACKNOWLEDGMENT

The authors would like to thank S. Dube for the fabrication and M. Thibault for the measurements.

TABLE III

<table>
<thead>
<tr>
<th>Frequency [GHz]</th>
<th>50% Error with δ'</th>
<th>Sidelobe level [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>88</td>
<td>Simulation</td>
<td>Measurement</td>
</tr>
<tr>
<td>55</td>
<td>0.11</td>
<td>0.10</td>
</tr>
<tr>
<td>60</td>
<td>0.13</td>
<td>0.12</td>
</tr>
<tr>
<td>65</td>
<td>0.17</td>
<td>0.18</td>
</tr>
<tr>
<td>70</td>
<td>0.19</td>
<td>0.20</td>
</tr>
<tr>
<td>75</td>
<td>0.22</td>
<td>0.24</td>
</tr>
<tr>
<td>80</td>
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TABLE IV

<table>
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<th>Antenna Gain at 88 GHz for Different Losses</th>
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<td>No dielectric, Metal loss only</td>
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</table>

Accurate Effective Permittivity Calculation of Printed Center-Fed Dipoles and Its Application to Quasi Yagi-Uda Antennas

A. Abbosh

Abstract—A closed-form method that accurately calculates the effective permittivity as seen by printed center-fed dipoles is presented. That accurate prediction enables the correct choice of the required dimensions of the dipole for a certain resonant frequency. For quasi Yagi-Uda antennas that usually have the driven element as a center-fed dipole, this means an accurate prediction of the required length of the driver, and subsequently all the other dimensions of the antenna without the need for extensive simulations and optimizations. The results obtained by full-wave electromagnetic simulations and measurements, over a wide range of design parameters, prove the validity of the derived method, which is based on the conformal mapping. The proposed method gives the correct estimation of the lengths of the dipoles with less than 10% error compared with the optimized design values. The conventional method results in estimations with as much as a 50% error.

Index Terms—Center-fed dipole, conformal mapping, printed dipole, quasi-Yagi antenna.

REFERENCES


I. INTRODUCTION

Printed center-fed dipoles are the main element of the many planar antennas widely used due to their compact size, low profile, light weight, low cost, and compatibility with other microwave devices [1]–[10]. A prime example, the quasi Yagi-Uda antenna, includes a driver, which is usually a center-fed dipole, reflector, and one or several directors.

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Color versions of one or more of the figures in this communication are available online at http://ieeexplore.ieee.org.
The required dimensions of antenna elements that use the center-fed dipole as a driver depend directly on the value of the effective relative permittivity as seen by the driver. For example, the length of the driven element of a quasi Yagi-Uda antenna is about half of the effective wavelength at the design frequency; the director is slightly shorter, whereas the reflector is slightly longer than the driver. The distance between the different elements of the antenna is around a quarter of the effective wavelength. The exact values of the director’s length and the distance between the antenna’s elements depend on the required radiation pattern.

An extensive review of the literature has been conducted to determine the method most often used to calculate the initial dimensions of center-fed dipoles and subsequently the other dimensions of antennas using that dipole. It has been found that the following formula [11] is usually adopted to estimate the effective relative permittivity \( \varepsilon_r \) when using a substrate with a relative permittivity \( \varepsilon_r \)

\[
\varepsilon_r = \frac{\varepsilon_r + 1}{2},
\]

(1)

The required driven dipole’s length \( l_d \) for a certain required resonant frequency \( f_r \) is

\[
l_d = \frac{c}{2 f_r \sqrt{\varepsilon_r}}.
\]

(2)

where \( c \) is the speed of light in free-space.

After examining the values of the initial and optimized dimensions of the driven dipole in many published antennas [1]–[10], it is observed that the optimized dimensions of the driven dipole is always significantly larger than the initial values calculated from (1) and (2). The large difference between the initial calculated values and the optimized values suggests that the estimation of the effective relative permittivity according to (1) is significantly larger than the real value. Actually, (1) is only correct when the utilized substrate that supports the dipole, and thus the antenna as a whole, is infinitely thick.

The work included in this communication presents a closed-form solution to that problem. The conformal mapping method used frequently to estimate the effective permittivity of transmission lines [12]–[14] is used to calculate the effective relative permittivity of the printed center-fed dipole. That permittivity can then be used to accurately calculate the required length of the driven element for a given resonant frequency and subsequently the overall dimensions of the antenna.

II. PROPOSED METHOD

An outline of a center-fed dipole is shown in Fig. 1. One of the arms of the dipole is labeled with \( \langle + \rangle \) whereas the other arm is labeled with \( \langle - \rangle \) to indicate the 180° phase difference between them. In order to find the equivalent relative permittivity as seen by the driven dipole, a quasi transverse electromagnetic propagation [12] is assumed for the structure of Fig. 1(a). Thus, the effective relative permittivity is determined from the effective capacitances of the driven element. The values of those capacitors depend on the distribution of the electric field lines between the two arms of the driven elements and with other parts of the antenna’s structure.

To understand the concentration of electric field lines in the structure of Fig. 1, the following assumptions are used. The two arms of the dipole are on opposite phase excitation. Moreover, the distance \( d \) between the driven dipole and the ground is more than one tenth of a wavelength and larger than the space \( S \) between the two arms of the dipole. Under those conditions, which are reasonable for antennas with center-fed dipoles, the electric field lines are concentrated between the two arms of the driven elements as in Fig. 1(b). Hence, the effective capacitance between the two arms of the dipole due to that field is the main parameter that defines the effective relative permittivity as seen by the driven element, and consequently, as seen by the antenna as a whole, if that driven element is not the only part of the antenna’s radiator.

From the electric field lines indicated in Fig. 1(b), it is possible to say that the total effective capacitance of the center-fed dipole includes three components. The first one is due to the field lines in free-space above the substrate; the second one is due to the field in the dielectric substrate, and the third one is due to the field at free-space below the substrate. With the help of the conformal mapping technique [12]–[14], the three components of the effective capacitance of the driven element can be calculated by doing several transformations so that each one of them can eventually be represented as a parallel-plate capacitance. The final result for the total effective capacitance \( C_e \) is

\[
C_e = \varepsilon_r \varepsilon_0 \frac{\pi}{2} K_1 + (\varepsilon_r - 1) K_2
\]

(3)

where \( \varepsilon_r \) is the relative permittivity of the substrate and \( w_d \) is the width of the dipole.

\[
\begin{align*}
K_1 &= \frac{K(k_1)}{K'(k_1)} \\
K_2 &= \frac{K(k_2)}{K'(k_2)}.
\end{align*}
\]

(4)

\[
\begin{align*}
k_1 &= \frac{l_d - S}{l_d + S} \\
k_2 &= \frac{\sinh \left( \frac{\pi(l_d - S)}{2h} \right)}{\sinh \left( \frac{\pi(l_d + S)}{2h} \right)}
\end{align*}
\]

(5)

\[
\begin{align*}
K(k) \quad \text{and} \quad K'(k) \quad \text{are the complete elliptical integral of the first kind} \quad \text{and} \quad \text{its complementary, respectively, of the parameter} \quad k.
\end{align*}
\]

From the conformal mapping transformations, the parameters \( k_1 \) and \( k_2 \) are related to the dimensions of the driven dipole and the substrate according to the following equation

\[
\begin{align*}
l_d &= \frac{l_d - S}{l_d + S} \\
S &= \frac{\sinh \left( \frac{\pi l_d}{2h} \right)}{\sinh \left( \frac{\pi l_d}{2h} \right)}
\end{align*}
\]

(6)

(7)

where \( l_d \) is the length of the dipole, \( S \) is the gap width between the two arms of the dipole, and \( h \) is the thickness of the substrate.
Under the practically used design limits (l_d > h and l_d \gg S), the parameters k_1 and k_2 can be approximated to

\[ k_1 = \frac{l_d}{l_d + 2S}, \quad (8) \]
\[ k_2 = e^{-\frac{\pi h}{l_d}}. \quad (9) \]

The effective relative permittivity, as seen by the driven dipole, is the ratio of the capacitor C_a to its value when \( \varepsilon_r = 1 \),

\[ \varepsilon_r = \frac{C_a}{C_{ea}}. \quad (10) \]

From (3) and (10)

\[ \varepsilon_r = 1 + \left( \frac{\varepsilon_r - 1}{2} \right) \frac{K_2}{K_1}. \quad (11) \]

Therefore, the resonant frequency \( f_r \) for a certain center-fed driven dipole of length \( l_d \) can be found from

\[ f_r = \frac{c}{2l_d \sqrt{\varepsilon_r}}. \quad (12) \]

It is to be noted, from (4)–(11), that the effective relative permittivity of the center-fed dipole depends on the substrate’s characteristics \( (\varepsilon_r, \varepsilon_s, \text{and } h) \), the gap width between the two arms of the dipole \( S \), and the dipole’s length \( l_d \). The effective relative permittivity, and thus, the required length for a certain resonant frequency do not depend on the width of the dipole \( w_d \) or the distance between the dipole and the ground \( d \). This final conclusion is the result of the practical assumption that \( d \gg S \). The existence of the ground at certain smaller distances from the dipole slightly changes the distribution of the electric field lines between the two arms of the director and thus, adds a certain small fringe capacitance to the effective total capacitance of the driven dipole. However, that capacitance is very small compared with the capacitance between the two arms of the dipole \( C_a \). The negligible effect of \( w_d \) and \( d \) are confirmed in the following section where the theory is compared with measurements.

It is possible to show from (8), (9), and (11) that, when the thickness of the substrate \( h \) is very large compared with the gap \( S \), \( \varepsilon_r \cong \frac{(\varepsilon_r + 1)}{2} \), which is equal to the estimated value by the conventional method (1). On the other hand, when \( h \) goes to zero, \( \varepsilon_r = 1 \), which is expected for an isolated dipole in free-space. Thus, it is sensible to say that, the conventional method is a reasonable tool for estimating the dimensions of printed dipoles only when built using very thin substrates with narrow gaps between the arms of the dipole.

The derived mode is used to calculate the effective relative permittivity \( \varepsilon_r \) of the center-fed dipole for a wide range of design parameters. The results are shown in Fig. 2. The effective relative permittivity calculated using the conventional method (1) is included in Fig. 2 for comparison.

The presented results indicate that increasing the gap width between the two arms of the dipole or decreasing the thickness of the substrate decreases \( \varepsilon_r \) significantly. Moreover, increasing the length of the dipole also decreases \( \varepsilon_r \). The effects of all those parameters are not included in the conventional method.

In the second tool, the presented method is used to verify the optimized lengths of center-fed dipoles that are used as part of developed

\[ \text{Fig. 2. Effective relative permittivity calculated using the proposed method for } \varepsilon_r = 4 \text{ compared with the conventional method.} \]

\[ \text{Fig. 3. Comparison between proposed method, conventional method, and full-wave simulations for } \varepsilon_r = 4 \text{ and } S = 1 \text{ mm. Indicated dimensions are in (mm).} \]

Therefore, it is expected that for antennas printed on thin high-permittivity substrates, the difference between the initial, as per the conventional method, and optimized lengths of the antenna elements, is quite large.

III. VALIDATION OF THE METHOD

To validate the accuracy of the derived method, different tools are adopted. First, the proposed method is used to calculate the resonant frequency of a center-fed dipole for a wide range of design parameters. The results of the calculations are compared with those obtained from full-wave electromagnetic simulations. A snapshot of the calculations is shown in Fig. 3. It is clear that the derived method is accurate in predicting the resonant frequency, whereas the conventional method gives a length estimation that is significantly less than the required value for a certain resonant frequency.

In order to validate the minor effect of the dipole’s width \( w_d \) on the resonant frequency, the results in Fig. 3 include two reasonable values for \( w_d \) (1 mm and 2 mm). It is observed that, increasing \( w_d \) by 100% only changes, and usually increases, the resonant frequency by less than 5% in the investigated cases.

In the second tool, the presented method is used to verify the optimized lengths of center-fed dipoles that are used as part of developed
antennas available in the literature [1]–[10] and an antenna developed in this work according to the design in [3]. To validate the proposed method when a ground plane and one or more parasitic elements exist as part of the antenna, the selected designs are all of the quasi Yagi-Uda antenna type.

The characteristics of the substrate and dipole from each design in [1]–[10] are used in (8)–(12) to calculate the required length of the dipole at the design frequency. For the papers that do not explicitly mention the design frequency ([2], [6], [8] and [9]), the first resonant frequency of the antenna is used in the calculations. Table I includes a comparison between the values obtained from the presented method, optimized values used to develop the antennas, and initial values from the conventional approach (1)–(2) when a thin substrate with high permittivity is used. The gap width are chosen to reveal the weakness of the conventional methods and the optimized value, is shown in the row D1 of Table I.

To cover other scenarios, the design presented in [3] is used to develop a quasi Yagi-Uda antenna that resonates at 10 GHz using the substrate Rogers RO6006 (εr = 6.2 and h = 0.54 mm) with S = 1 mm and w0 = 0.4 mm. The substrate’s characteristics and the gap width are chosen to reveal the weakness of the conventional approach (1)–(2) when a thin substrate with high permittivity is used. The antenna is designed using (8)–(12) for a resonant frequency of 10 GHz. A comparison, between the initial value using the presented and conventional methods and the optimized value, is shown in the row D1 of Table I.

It is clear from Table I, that the proposed method is successful in estimating the required dipole’s length with less than 10% error in all the listed cases, which include a wide range of substrates, resonant frequencies, ground dimensions and shapes, and number and position of parasitic elements. The method is successful in the accurate prediction of the effective relative permittivity, and thus, the dipole’s length, even when the dipole is designed as a meandered structure [9] or dual meandered structure for multibands [10]. The accuracy of the predictions listed in Table I validates the suggested method and confirms the principle utilized in this communication that says the effective relative permittivity, as seen by the driven element, is the effective relative permittivity of the antenna as a whole. The ground plane and parasitic elements have limited effect on that value assuming that they are not very close to the driven element.

The conventional approach fails to accurately predict the required dimensions in all the listed cases in Table I. The difference between the lengths calculated using the conventional approach and the optimized values is significant (30%–70%) in the listed cases. The only design where the conventional method gives a reasonable estimation for the required length of the dipole is from [7], which uses a relatively thick substrate with low permittivity, as expected in the derived theory.

### IV. Conclusion

A closed-form method for accurately calculating the effective relative permittivity of printed center-fed dipoles is presented. The proposed method enables the correct choice of the required dimensions of the dipole for a certain resonant frequency. For quasi Yagi-Uda antennas, for example, this means an accurate prediction of all the dimensions of the antenna.

### REFERENCES


### Table I

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<th>Proposed Method</th>
<th>Optimized (Design) Value</th>
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**Notes:**
- The COMPARISON BETWEEN THE DIPOLE’S LENGTH (mm) ESTIMATION USING THE CONVENTIONAL METHOD, PROPOSED METHOD, AND OPTIMIZED (DESIGN) VALUE
- The values in parentheses correspond to the conventional method.
- The values in bold correspond to the proposed method.
- The values in italic correspond to the optimized (design) value.
Ultra-Wideband Quasi-Yagi Antenna Using Dual-Resonant Driver and Integrated Balun of Stepped Impedance Coupled Structure

A. Abbosh, Senior Member, IEEE

Abstract—A quasi-Yagi antenna that has an ultra-wideband performance is presented. To enable that performance, the antenna utilizes a dual-resonant driver and a balun formed using a stepped-impedance coupled structure. The driver is designed to be dual-resonant by loading it with an inductor in the form of a short section of narrow microstrip line at a certain position. The balun includes a T-junction of microstrip lines and two pairs of stepped-impedance coupled lines. The simulated and measured performance of the integrated antenna indicates less than -10 dB reflection coefficient, 3.6-4.5 dBi gain, 13-17 dB front-to-back ratio, less than -19 dB cross-polarization and more than 90% efficiency across more than 75% fractional bandwidth centered at 7.5 GHz.

Index Terms—Quasi-Yagi antenna, balun, wideband antenna, planar antenna.

I. INTRODUCTION

Quasi-Yagi antennas are widely used for many applications due to their compact size, low profile, lightweight, low-cost, and compatibility with microwave circuits. This type of antenna is known to exhibit moderate gain, endfire radiation pattern with modest bandwidth. It consists of three main components; driver, reflector, and one or several parasitic directors. The truncated microstrip ground plane of the antenna acts as a reflector for the transverse-electric surface wave generated by the driver. The parasitic directors are used to enhance the radiation in the forward endfire direction.

Since the driver element of the planar quasi-Yagi antenna is a balanced structure, whereas the feeder is usually an unbalanced transmission line, most of the research work in the literature concerning quasi-Yagi antenna focuses on designing different types of balanced-to-unbalanced (balun) feeding structures. The various techniques used in the design of baluns and other parts of the antenna aim at broadening the impedance bandwidth of the antenna [1]-[12].

In [1] and [2], a coplanar waveguide (CPW)-to-slotline transition is used to build a quasi-Yagi antenna. In another approach, a planar quasi-Yagi antenna utilizing a microstrip-to-coplanar stripline (CPS) balun is proposed in [3], [4]. The combination of truncated ground plane and a feeding structure in the form of two parallel strips on the two layers of the substrate is proposed [5]. A similar approach, but in this case using a uniplanar structure fed by a CPW, is adopted in [6] and [7]. In [8], a four-port bandpass filter is converted into a three-port balun that feeds a quasi-Yagi antenna. The traditional coaxial balun is converted into a planar balun and used to feed a quasi-Yagi antenna in [9] and [10].

In another method, a combination of microstrip to CPS transition and artificial transmission line is used for the feeder, whereas the driver is designed in the form of meandered microstrip line [11]. In a different configuration, a partial ground plane that is separate from the reflector and a notched microstrip patch as driver are used [12].

By inspecting the results presented in [1]-[12], it is clear that the available quasi-Yagi antennas achieve at most around 50% fractional bandwidth assuming the -10 dB reflection coefficient as a reference. In this paper, a quasi-Yagi antenna with more than 75% fractional bandwidth and 13 dB front-to-back ratio across is presented. The main features of the proposed antenna are the ultra-wideband (UWB) balun and a dual-resonant driver.

The utilized balun in this work is based on the combination of stepped-impedance structures of T-junction and parallel-coupled lines. The driver of the antenna is designed to be a dual-resonant dipole by loading it by an inductor in the form of a short section of narrow microstrip line at a certain distance from the feed point. The design is validated via theory, simulations and measurements. The developed antenna is designed to operate across a band centered at 7.5 GHz. The choice of this band is related to the future use of the antenna in a wideband microwave imaging system.

II. PROPOSED ANTENNA

The configuration of the proposed UWB quasi-Yagi antenna is depicted in Fig. 1. The design in this work includes two main contributions. The first one is the ultra-wideband balun that includes a stepped-impedance T-junction of microstrip lines and two pairs of stepped-impedance parallel-coupled microstrip lines. The two pairs are arranged as indicated in Fig. 1 to ensure the required odd-mode excitation to the center-fed driver of the antenna across an ultrawideband. The odd-mode excitation here means that the two signals supplied to the two arms of the driver are out-of-phase with each other. The second contribution is the dual-resonant driver needed for the UWB performance. The dual-resonant is
accomplished by loading the driver by a suitable inductor at a
certain location to ensure that the resonant frequencies of the
driver are properly positioned for wideband coverage. The
inductor is implemented in the form of a short section of
narrow microstrip line.

A. Balun design

As depicted in Fig. 1, the utilized balun includes two main
parts. The first one is the stepped-impedance T-junction of
microstrip lines needed to divide the input signal from Port#1
into two equal parts. Those two signals are supplied to the
input terminals of two pairs of stepped-impedance parallel-
coupled lines, which represent the second main part of the
balun. Each pair of those coupled lines has four terminals.
Two of them represent the input and output terminals of the
coupled structure, whereas the remaining two terminals are
connected to the ground using short circuit vias in one pair and
left as open-ended microstrip lines in the other pair. This
arrangement guarantees producing two out-of-phase signals at
the two output ports of the balun (Port#2 and Port#3). The two
output ports of the balun are connected with the coplanar
stripline (CPS) needed to feed the driver of the antenna.

The T-junction has a stepped impedance input line that is
used to improve the matching with the 50 Ω input port
(Port#1). The two output terminals of the T-junction are
tapered for an improved matching with the two input terminals
of the coupled structures. The coupled structures are also
designed in the form of stepped impedance coupled lines for
ultra-wideband performance [13]. To that end, each of the two
coupled structures includes three sections as depicted in Fig. 1.
The two side sections are similar with a length of \( l_{c1} \), whereas
the central section has a length \( l_{c2} \). Based on the dimensions of
the coupled structures, the side sections have an
electromagnetic coupling factor \( c_1 \), whereas the central section
has a coupling factor \( c_2 \).

To find the required values for \( l_{c1}, l_{c2}, c_1, \) and \( c_2 \) for an
ultra-wideband performance, the even-odd mode analysis for
four-port devices [13] are utilized. Assuming a perfect
matching at the three ports of the balun, it is possible to show
that the proposed balun can cover the band from 3 GHz to 12
GHz with 180°±10° differential phase between the two
balanced outputs and less than 0.5 dB insertion loss using
\( c_2 = 2c_1 = 0.8 \) and \( l_{c2} = 2l_{c1} = \lambda_c/6, \lambda_c \): The effective
wavelength at the center frequency of the band \( f_c = (3+12)/2 = 7.5 \) GHz. This solution means that the
coupled structure in the balun should have a total length
\( (2l_{c1} + l_{c2}) \) that is equal to one third of the effective
wavelength at the center of the band.

The value of the required coupling shows that for a UWB
performance, the side sections are loosely coupled. From the
fabrication perspective when using the printed circuit
technology, it is easy to realize the required loose coupling of
0.4 at the two side-coupled sections using conventional
parallel-coupled microstrip lines.

### Table I Design Parameters (Dimensions in mm and Capacitor in pF) of the Final Design

<table>
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As per the calculated design values, the central section of
the coupled structure should be tightly coupled. From the
mode analysis terminology, the tight coupling requires low
odd-mode impedance and high even-mode impedance.
Therefore, a chip capacitor is connected between the two
coupled lines at the middle of each of the central coupled
sections to increase its odd-mode capacitor, and thus, to
decrease its odd-mode impedance without the need for a
narrow gap [14]. Moreover, a slotted ground is created
underneath each of the central coupled sections to reduce the
even-mode capacitor, and thus, increase the even-mode
impedance, while using an easy-to-manufacture coupled
structure [15].

From the required coupling factors, it is possible to estimate
the physical dimension of the coupled structure and the
capacitance value of the chip capacitors following the
conformal mapping analysis [16].

B. Radiator design

As shown in Fig. 1, the designed balun is integrated with a
quasi-Yagi antenna that includes one driver, and one director.
The ground plane of the balun operates as a reflector. Since
the designed balun is proven to have an ultra-wideband
performance, it is expected that the bottleneck in the way to
realize an UWB performance of the whole integrated antenna
is the driver.

It is well known that since the driver of the antenna is a half
wavelength resonator, the band is usually limited. To avoid
this bottleneck, the driver is designed to be dual-resonant
structure. The proposed driver is depicted in Fig. 1. The driver
of total length \( l_9 \) is designed to resonate close to the lower
The frequency of the targeted band, whereas the part of the driver of length $l_4$ is designed to resonate at the middle of the band. To enable that dual-resonant behavior, the driver is loaded with an inductor in the form of a short section of narrow microstrip line at the boundaries between $l_3$ and $l_4$ as illustrated in Fig. 1.

At the lower part of the frequency band, the inductor has a low impedance, and thus, the effective radiator is the dipole of total length $l_3$ which is chosen so that the dipole resonates at a low frequency $f_1$. As the frequency increases, the inductor's impedance increases in the value till it becomes a virtual open circuit. Thus, the effective radiator is the dipole of length $l_4$. The length $l_4$ is chosen so that this short dipole resonates at the center of the targeted band $f_2$. By the proper choice of the frequencies ($f_1$ and $f_2$) and the dimensions of the narrow microstrip that represents the inductor, the band of the integrated antenna can be increased significantly. The dimensions of the microstrip line that forms the inductor can be calculated according to the procedure in [17].

To choose suitable lengths for the dual-resonant driver, the effective dielectric constant as seen by the driver is needed. The recently developed theory for the accurate estimation of the effective dielectric constant as seen by center-fed dipoles is utilized [18]. According to that theory, the effective dielectric constant as seen by the driver is

$$\varepsilon_r = 1 + 0.5(\varepsilon_r - 1)K_2/K_1$$

(1)

$$K_1 = K(k_1)/K'(k_1)$$

(2)

$$K_2 = K(k_2)/K'(k_2)$$

(3)

$$\varepsilon_r:$$ dielectric constant of the substrate, $K(k_1)$ and $K'(k_1)$: the complete elliptical integral of the first kind and its complementary, respectively, of the parameter $k_1$. The parameters $k_1$ and $k_2$ for a certain director's length $l_1$, gap between the two arms of the driver $S_{d'}$, and thickness of the utilized substrate $h$ are given as [18]

$$k_1 = c\sqrt{\varepsilon_r}$$

(4)

$$k_2 = c\sqrt{\varepsilon_r}/2h$$

(5)

Therefore, the resonant frequency $f_r$ for a driver's length $l_1$ can be found from

$$f_r = c/[2l_1(\varepsilon_r)]$$

(6)

$4\pi \times 10^8$ m/s.

In the current antenna, the driver is designed to have the two resonant frequencies, $f_1=5$ GHz and $f_2=7.5$ GHz. Thus, the lengths $l_2$ and $l_4$ are calculated from (1)-(6) to be 21.3 mm and 13.6 mm, respectively, assuming the substrate Rogers RT6010 ($\varepsilon_r =10.2$, $h =0.64$ mm). In the calculation, the gap $S_{d'}$ is assumed to be equal to the gap between the two output ports of the balun (2 mm) for a direct and easy connection between the output ports of the balun and the CPS feeder of the driver.

Concerning the director, it is designed to have a length $l_6$ equal to $0.25\lambda_c$, whereas the distances between the driver and the ground plane $d_2$, and between the driver and the director $d_3$ are chosen to be $0.2\lambda_c$ and $0.15\lambda_c$, respectively, as per the guidelines of Yagi antenna design. The wavelength ($\lambda_c$) at the center of the band (7.5 GHz) can be calculated from the effective dielectric constant (1). Using (6), $l_6$, $d_2$, and $d_3$ can be found to be 6.8 mm, 5.4 mm, and 4 mm, respectively.

The final dimensions are then obtained using the sequential nonlinear programming in the software HFSS. The whole structure (antenna and balun) are optimized for the widest possible -10 dB reflection coefficient bandwidth that is centered at 7.5 GHz, and more than 10 dB front-to-back ratio. The final dimensions in (mm) using the substrate Rogers RT6010 ($\varepsilon_r =10.2$, $h =0.64$ mm) are listed in Table I. The overall dimensions of the antenna are 27 mm×32 mm.

III. RESULTS AND DISCUSSIONS

The designed antenna is fabricated (Fig. 2) and tested. The simulated and measured reflection coefficients of the antenna are presented in Fig. 3. The antenna covers the band from 4.7 GHz to 10.4 GHz, which is equivalent to 75% fractional bandwidth, with less than -10 dB reflection coefficient. The simulated and measured results are in good agreement across the investigated band. The driver of the antenna is designed to have the first two resonant frequencies at 5 GHz and 7.5 GHz. The simulated results indicate that those two resonances appear at 5.1 GHz and 7.3 GHz, whereas the measured results show the two resonances at 5.1 GHz and 7.5 GHz. Those results validate the procedure used in designing the driver.

To validate the directive properties of the antenna, the gain is simulated and measured. The simulated values shown in Fig. 4 indicate that the gain varies between 4 dBi and 5 dBi across the band from 4.7 GHz to 10.4 GHz. Across the same band, the measured gain varies between 3.6 dBi and 4.5 dBi. Apart from the good agreement at the center of the investigated band, the measured results indicate generally slightly lower values for the gain than the simulated values.

The radiation pattern of the antenna is measured at the two principal planes (H-plane and E-plane) at three frequencies (5 GHz, 7.5 GHz, and 10 GHz). The results depicted in Fig. 5 confirm the directive properties of the antenna. The front-to-back ratio of the antenna varies from 13 dB at 5 GHz to 17 dB at 10 GHz. At the endfire direction of the pattern, the copolarized signal is higher than the cross-polarized signal by more than 19 dB at the three selected frequencies.

To evaluate the effect of the different dielectric and conductive losses on the radiation of the antenna, the radiation efficiency is calculated using HFSS as depicted in Fig. 4. It is revealed that the efficiency is more than 93% across the band from 4.5 GHz to 10.5 GHz. To verify the simulated values, the radiation efficiency is also estimated using the measured 3 dB beamwidths and gains [19]. It is found that the efficiency is more than 90% across the band from 4.5 GHz to 10.5 GHz.

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Fig. 2 Top (a) and bottom (b) views of the developed antenna.
Fig. 3 Reflection coefficient of the designed antenna.

Fig. 4 Gain and efficiency of the antenna. Solid lines: Simulations.

Fig. 5 Measured radiation pattern of the antenna.

IV. CONCLUSION

An ultra-wideband quasi-Yagi antenna has been presented. The two main features of the presented antenna are the balun and the dual-resonant driver. The balun includes stepped-impedance microstrip lines forming a T-junction and two pairs of coupled lines. The driver is designed to have two main resonances by loading it at a certain position by an inductor of short section of narrow microstrip line. The simulated and measured performance of the integrated antenna indicate less than -10 dB reflection coefficient, 3.6-4.5 dBi gain, 13-17 dB front-to-back ratio, less than -19 dB cross-polarization and more than 90% efficiency across 75 % fractional bandwidth.

REFERENCES

A Planar UWB Antenna With Signal Rejection Capability in the 4–6 GHz Band

A. M. Abbosh, M. E. Bialkowski, Fellow, IEEE, J. Mazierska, and M. V. Jacob

Abstract—The design of a compact planar antenna featuring ultra wideband performance and simultaneous signal rejection in the 4–6 GHz band, assigned for IEEE802.11a and HIPERLAN2, is presented. The design is demonstrated assuming RT6010LM substrate with a relative dielectric constant of 10.2 and thickness of 0.64 mm. The presented results show that the designed antenna of 27 mm × 20 mm dimensions has a bandwidth from 2.7 GHz to more than 10 GHz excluding the rejection band. The antenna features near omnidirectional characteristics and good radiation efficiency.

Index Terms—Planar antenna, slot antenna, ultra wideband (UWB) antenna.

I. INTRODUCTION

R ECENT years have witnessed an increased interest in ultra wideband (UWB) antennas since the adoption of UWB technology by the U.S.-FCC in 2002 [1]. Parallel to this recent interest in UWB antennas, research has also focused on a multilayer dielectric substrate approach to the design of front end modules to reduce the size of wireless transceivers [2]. In order to achieve integration between the radio frequency (RF) front end circuitry and a radiating structure, an UWB antenna should preferably be of planar format. Several planar UWB antenna designs, which have the potential to meet such requirements, were reported in [3]–[7].

As UWB systems need to meet the condition of harmonious operation with some already existing standards such as the IEEE802.11a and HIPERLAN2A, rejecting a particular sub-band within the 4.0–6.0 GHz band becomes an important design requirement. One method to meet the new constraint is to use a filter rejecting undesired frequencies. However, a more attractive solution is to employ an UWB antenna with sub-band rejection capabilities, as shown in [8]. One problem with the design presented in [8] is that it is nonplanar and thus does not fulfill the requirement for integration with a multilayer dielectric RF front end circuitry.

In this article, a fully planar UWB antenna with a narrow band rejection capability within the 4–6 GHz band is described. A simple design technique is introduced to accomplish this goal.

Fig. 1. Configuration of the UWB antenna with capability of rejecting a narrow sub-band: (a) a three-dimensional (3-D) model with the feeding structure and (b) layout showing different design parameters.

Section II of this letter describes the proposed antenna design. Computer simulations and measurements concerning voltage standing wave ratio (VSWR), radiation pattern, and gain of the proposed antenna are shown in Section III. Section IV concludes the article.

II. DESIGN

The configuration of a planar UWB antenna with the capability of rejecting frequencies within the 4.0–6.0 GHz band is illustrated in Fig. 1.

The antenna structure is assumed to be in the $x'y'$ plane with its higher dimension extending along the $y'$-axis. The radiating slot is the result of intersecting of two circles in a conductive layer on top of the substrate. The antenna is fed with a coplanar waveguide (CPW). The observed transition from CPW to a coaxial probe can be regarded as via in a multilayer front end module. A tuning slot to reject a signal within the 4–6 GHz band is introduced in the second circle. Steps used to design this antenna are summarized as follows.

1) Depending on the lowest frequency of operation ($f_1$), thickness of the substrate ($h$) and its dielectric constant ($\varepsilon_r$), width ($i'$), and length ($l$) of the antenna structure are calculated from [9] as

$$i' = \frac{c}{2f_1} \sqrt{\frac{2}{\varepsilon_r + 1}}$$

$$l = \frac{c}{2f_1 \sqrt{\varepsilon_{RE}}} - 2\Delta$$

$$\varepsilon_{RE} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ 1 + \frac{L}{i'} \right]^{-0.5}$$
\[ \Delta l = \frac{1}{2} \left( \frac{\varepsilon_{re} + \varepsilon_{im}}{\varepsilon_{re} - \varepsilon_{im}} \right) \left( \frac{4\pi l_1}{\varepsilon_{re} + \varepsilon_{im}} \right) \]  

(4)

where \( c \) is the speed of light in free space and \( \varepsilon_{re} \) is the effective dielectric constant. In the undertaken approach, the parameter \( \Delta l \) is neglected as it usually takes a small value in comparison with \( l \).

2) The radiating slot is formed by cutting a circle (Circle1) from the ground plane designed above and adding a second circle (Circle2) to the ground plane in the manner shown in Fig. 1. Diameters of Circle1 and Circle2 (\( \overline{P}_1 \) and \( \overline{P}_2 \)), in the preliminary stage of the design are equal to

\[ \overline{P}_1 = \overline{P}_2 = \overline{P}_1 \cdot \frac{1}{2} \]  

(5)

Usually, \( \overline{P}_1 \) is taken to be lower than the value mentioned in (5) to allow some space for the ground plane around the radiating slot.

3) Centers of the two circles are shifted by \( \overline{x}_1 \) and \( \overline{x}_2 \) from centre of the ground plane. Parts of the two circles that extend outside the ground plane from one direction are cut to form the CPW needed to feed the antenna. Values of \( \overline{x}_1 \) and \( \overline{x}_2 \) are chosen in order to maintain the required impedance bandwidth.

4) In order to tune out the undesired band a tuning slot is made in the second circle. The slot is in the shape of half circle with diameter \( \overline{P}_3 \) chosen according to the following equation:

\[ \overline{P}_3 = \frac{c}{2 f_{in} \sqrt{\varepsilon_{re}}} \]  

(6)

where \( f_{in} \) is centre of the undesired band. Position of the slot and its width are selected to control the rejected band.

Design parameters (\( \overline{P}_1, \overline{P}_2, \overline{P}_3, \overline{x}_1, \) and \( \overline{x}_2 \)) for the UWB slot antenna obtained using this method are optimized with Ansoft HFSSv2 and an in-house developed program written in Microsoft-Perl 5.6 [10]. The initial and optimized values assuming \( f_1 \) of 2.5 GHz and \( f_{in} \) of 5.4 GHz are given in Table I. The optimization was performed for the best impedance bandwidth excluding the undesired band. The optimized values differ by less than 10% from the original ones. This result gives a lot of confidence in the presented design formulas.

### III. RESULTS AND DISCUSSION

The UWB antenna with signal rejection capabilities in the 4–6 GHz band was designed assuming RT6010LM substrate with a dielectric constant equal to 10.2, tangent loss \( t\tan h = 0.0023 \) and thickness of 0.64 mm, which was readily available to the authors.

Fig. 2 shows the simulated and measured results for the VSWR with frequency for the designed antenna (with dimensions presented in Table I) without and with the tuning slot.

For the antenna without the tuning slot both the computed and measured VSWR characteristics reveal UWB behavior with bandwidth from 2.7 GHz to more than 10 GHz assuming VSWR = 2 as a reference. When the tuning slot is included, the worst VSWR occurs over a narrow sub-band within the 4–6 GHz. The choice is dependent on the length of the tuning slot. In the pass-band the VSWR characteristics of the original UWB antenna are only slightly affected by the presence of the tuning slot. Effect of using tuning slot with different diameters is shown. It is clear that position of the rejected band can be controlled by adjusting diameter of the slot. Increasing diameter of the tuning slot shifts the rejected band to a lower value by a proportion which is compatible with the design (6). It has been observed that in order to maintain the behavior of the antenna within the desired band the center of the tuning slot should coincide exactly with center of Circle2. The rejected bandwidth can be increased using a wider tuning slot. This is confirmed by both simulated and measured results, which show only small discrepancies.

From the UWB applications point of view the antenna is required to have an omnidirectional radiation pattern. This requirement is fulfilled by the designed antenna, as shown by simulated results in Fig. 3.

The designed antenna measured gain is presented in Fig. 4. This figure reveals that the gain (in dB) increases approximately linearly with frequency for the band 3 GHz until 10 GHz and is between 0.4 to 4.6 dB for the original antenna (without the tuning slot) and for the modified antennas (with the tuning slot) outside the rejected band. The gain in the rejected band is found

---

**TABLE I**

<table>
<thead>
<tr>
<th>Design Parameters</th>
<th>Designed value (mm)</th>
<th>Experimental value (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( M^* )</td>
<td>12.5</td>
<td>25.0</td>
</tr>
<tr>
<td>( l )</td>
<td>125</td>
<td>25</td>
</tr>
<tr>
<td>( D_1 )</td>
<td>12.5</td>
<td>12.6</td>
</tr>
<tr>
<td>( D_2 )</td>
<td>8.9</td>
<td>9</td>
</tr>
<tr>
<td>( y_1 )</td>
<td></td>
<td>3</td>
</tr>
<tr>
<td>( y_2 )</td>
<td></td>
<td>6.8</td>
</tr>
</tbody>
</table>

**Fig. 2.** Measured and simulated VSWR against frequency for the designed UWB antenna.
to be as low as 4.5 dB. The HFSS simulations have shown that this antenna features more than 90% radiation efficiency in the pass-band.

IV. CONCLUSION

In this letter, a simple design method for a planar UWB antenna with signal rejection capabilities over the 4–6 GHz sub-band has been presented. The proposed radiating element is a slot formed by the intersection of two circles which also include the CPW as a feed. A tuning slot is responsible for the sub-band rejection. Computer simulations and measurements have proved the validity of simple design formulas. The designed UWB antenna features omnidirectional radiation pattern and good radiation efficiency. The presented antenna configuration and its design method should be of considerable interest to the designers of UWB front ends employing multilayer substrates.

REFERENCES

pression for a given substrate condition. The stop-band level is also lowest for the proposed EBG power plane.

Since the EBG power plane is also used as a reference plane for signal lines, there could be a signal integrity problem due to the many slots on the reference plane. The possible signal integrity problem can be resolved using differential signaling, embedded EBG power plane [6], or stitching capacitors [7].

4. CONCLUSION

A new EBG power plane comprised of thin spiral strip lines and hybrid-cells is proposed for ultra-wideband suppression of the GBN propagation. The hybrid-cell EBG structure offers both of a high series inductance and a shorter period of the EBG lattice. More than 30 dB GBN suppression has been measured in a wide range of frequencies from 370 MHz to 8.27 GHz. We expect the hybrid-cell EBG power plane be applied for high speed and mixed modes PCB design by the wideband GBN suppression.

REFERENCES


DESIGN OF UWB PLANAR ANTENNA WITH SUBBAND REJECTION CAPABILITY

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ABSTRACT: A simple guideline for designing a compact ultra wideband planar antenna is presented. The proposed antenna has the capability to reject any undesired subband within the whole bandwidth of its operation by incorporating a tuning slot within the radiating structure. Using GML1032 dielectric substrate with a relative permittivity of 3.2, the designed antenna features a compact size of $27 \times 27$ mm$^2$. Results of the measurement show that the designed antenna has a 10 dB return loss bandwidth from 2.8 GHz to more than 11 GHz excluding any undesired subband, such as the 4.9–5.9 GHz band assigned for IEEE802.11a and HIPERLAN/2. The antenna has near omnidirectional characteristics, and its radiation efficiency is higher than 90% over the whole pass-band. © 2007 Wiley Periodicals, Inc.


Key words: planar antenna; subband rejection; ultra wideband antenna

1. INTRODUCTION

Ultra Wideband (UWB) is a short range communications technology, which transmits the information in the form of very narrow pulses. This former military technology has gained a lot of popularity among researchers and wireless industry after the FCC permitted the marketing and operation of UWB within the range of 3.1–10.6 GHz [1]. The global interest in the UWB technology is increasing very fast because of the capability of this license exempt wide bandwidth system to yield low cost, low energy, short range, and extremely high capacity wireless communication links.
Compact and cheap UWB antennas are needed for the numerous UWB applications, such as wireless communications, indoor positioning, and medical imaging. The printed planar monopole antennas are usually the preferred candidates as they can be easily fabricated, and are of low cost and light weight. The biggest challenge to fit these antennas in a UWB system is to widen their impedance bandwidth while maintaining high radiation efficiency.

Many UWB antennas have been reported in the literature, such as the planar volcano-smoke slot antenna [2, 3], the planar triangular monopole [4, 5], the multistructure coplanar waveguide fed [6], the circular disc monopole [7], the stepped patch [8], and the notched rectangular patch [9]. The main drawback for the above mentioned designs is that they use a trial and error method with the help of a simulation tool to get the desired UWB operation.

UWB antennas should also have the capability to reject any interference with existing wireless networking technologies such as the subband 4.9–5.9 GHz for IEEE 802.11a in the USA, and HIPERLAN/2 in Europe [10]. This is due to the fact that UWB transmitters should not cause any electromagnetic interference on nearby communication systems such as wireless LAN (WLAN) applications. Therefore, UWB antennas with notched characteristics in WLAN frequency band are desired. Some designs have appeared recently for UWB antennas with subband rejection capability [11, 12]. The design of the antennas in those articles depends on the trial and error method with the help of sophisticated software to achieve the required performance.

This article describes a design method for a compact planar UWB antenna fed by a coplanar waveguide (CPW). The presented method also shows how to design a tuning slot to reject any undesired subband within the UWB frequency spectrum. The method presented in this article is used to design an UWB antenna that rejects the 4.9–5.9 GHz band, which is assigned for the IEEE802.11a and HIPERLAN/2. Results of measurements for two antennas, designed using the proposed method, are presented in order to prove the validity of the proposed method and to show the behavior of the return loss, radiation efficiency, and gain of the manufactured antennas.

2. DESIGN

The configuration of the proposed ultra wideband (UWB) antenna with the capability of rejecting frequencies over a certain subband is illustrated in Figure 1. The radiating structure is formed by the connection of half an ellipse with a rectangular patch. The antenna is assumed to be fed using a coplanar waveguide (CPW) to enhance its broadband characteristics. The ground plane, which is located around the CPW feeder, is in a shape of half an ellipse. The ground plane results in a horizontal elliptical dipole of a length equal to a half wavelength, and it is elevated from the ground by a distance which is about a quarter wavelength.

The design procedure starts by finding dimension of the antenna feeder to give 50 Ω characteristic impedance \( Z_0 \). This can be achieved using the following equations [13]:

\[
Z_0 = \frac{30\pi K'(k)}{\sqrt{k' K(k)}}, \quad (1)
\]

\[
k = \frac{s}{s + 2w_s}, \quad (2)
\]

\[
e_v = \frac{e + 1}{2} - \frac{A + B}{2}, \quad (3)
\]

\[
A = \tan h[1.785 \log(h/w_s) + 1.75], \quad (4)
\]

\[
B = \frac{k w_s}{h}[0.04 - 0.7k + 0.01(1 - 0.1e)[0.25 + k] - \frac{1}{2} s], \quad (5)
\]

where \( K(k) \) is the first kind elliptical integral and \( K'(k) = K(\sqrt{1-k^2}) \), \( s \) is the central conductor width, whereas, \( w_s \) is the slot width of the CPW, \( h \) is the substrate thickness, and \( e \) is the dielectric constant of the substrate.

Figure 1 Configuration of the proposed UWB antenna. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

Depending on the lowest frequency of operation \( f_1 = 3.1 \) GHz, thickness of the substrate and its dielectric constant, width \( w \) and length \( l \) of the antenna structure are calculated as

\[
w = l = \frac{c}{2f_1\sqrt{e}} \quad (6)
\]

where \( c \) is speed of light. The antenna is assumed to be in the \( xy \) plane, with the dimension \( w \) extending along the \( x \) axis. Note that the length and width of the antenna structure, according to (6), are equal to half of the wavelength inside the substrate medium.

The ground plane of the antenna, which is located around the CPW, is half of an ellipse with major diameter equal to \( w \). The secondary diameter of the ground plane can be chosen to be within the range of \((0.5w)\) to \((0.75w)\). The radiating structure, which is located at a distance \( g \) from the end of the ground plane, consists of two parts. The first part, which is connected directly with the feeder, is half an ellipse with the same dimensions as for the ground plane. The second part is a rectangular patch, which extends from the end of the half ellipse till the end of the substrate. Parametric analysis on the best value for \( g \) indicates that it should be less than thickness of the substrate \( h \) to get the widest bandwidth. In the design procedure of this article, \( g \) is assumed to be equal to \((h/3)\).

Note that the above choice for dimension of the radiator and the ground plane results in a horizontal elliptical dipole of a length equal to a half wavelength, and it is elevated from the ground by a distance which is about a quarter wavelength.

The above design procedure results in an antenna which covers the whole UWB range from 3.1 to 10.6 GHz. If it is required to
To reject a certain subband within that range, an elliptical tuning slot with length \( l_s \) and width \( w_s \) can be incorporated in the radiating structure at a distance \( p_s \) from the feeding point, see Figure 1. The required length of the slot \( l_s \) and its position \( p_s \) depend on value of the center frequency of the undesired subband \( f_c \) and the dielectric constant of the substrate according to the following equations.

\[
\begin{align*}
  l_s &= \frac{c}{2f_c\sqrt{\varepsilon_r}} \\
  p_s &= \frac{c}{4f_c\sqrt{\varepsilon_r}}
\end{align*}
\]

To maintain the performance of the antenna across the desired band, width of the tuning slot should be chosen such that

\[
w_s \leq \frac{c}{f_h\sqrt{\varepsilon_r}}
\]

where \( f_h = 10.6 \text{ GHz} \) is the highest frequency of operation.

### 3. RESULTS

The proposed UWB antenna was designed using GML1032 substrate with a dielectric constant equal to 3.2, tangent loss \( \tan \delta = 0.004 \), and thickness of 1.52 mm. Values of the design parameters shown in Figure 1 calculated using \((1–6)\) are: \( w = l = 27 \text{ mm}, s = 2.8 \text{ mm}, w_c = 0.23 \text{ mm}, g = 0.5 \text{ mm} \).

Concerning the tuning slot, assume that it is required to design the antenna to reject the subband 4.9–5.9 GHz from its response. The center frequency of the undesired subband \( f_c \) is equal to \((4.9 + 5.9)/2 = 5.4 \text{ GHz} \). Therefore, dimensions of the slot \((l_s, p_s, \text{ and } w_s)\) can be calculated, using \((7–9)\), to be 15.6 mm, 7.8 mm, and 1 mm, respectively.

According to (7), if it is required to tune out a different subband with different center frequency then it is possible to use the same antenna structure with dimensions calculated using \((1–6)\), but with different slot parameters. In Figure 2 variation of the required slot length and position with center frequency of the rejected subband are shown together with the optimum values obtained using the software HFSS. For all the cases shown in Figure 2, slot width \((w_s)\) was constant at 1 mm. It is clear from Figure 2 that \((7)\) and \((8)\) give an accurate estimation for dimensions of the required slot.

Two samples of the proposed antenna were manufactured, one without a tuning slot and the other with a tuning slot to reject the subband 4.9–5.9 GHz. Characteristics of the developed antennas were tested using a vector network analyzer in an anechoic chamber. Figure 3 shows variation of the return loss with frequency for the developed antennas. The measured characteristics of the antenna without a slot reveal UWB behavior with bandwidth from 2.8 GHz to more than 11 GHz assuming a 10 dB return loss reference. It is also clear from Figure 3 that, for the antenna with a tuning slot, the undesired subband was tuned out, whereas the wideband behavior of the antenna was maintained.

From the UWB applications point of view, the antenna is usually required to have an omnidirectional radiation. Concerning the designed antenna, this requirement is fulfilled over the whole bandwidth as shown in Figure 4, especially at the \(yz\) plane. The measurements reveal that, at the high frequency band, the main beam tilts a little in the forward direction, as in the radiation pattern of Figure 4 at 6 and 9 GHz for the \(yz\) plane. The tilt in the radiation pattern is common for many wideband planar antennas that have been reported in many recent papers. This is because of the excitation of higher order current distributions (modes) on the antenna structure at higher frequencies.

The measured radiation patterns of the antenna with a slot are similar to the results shown in Figure 4 and therefore, they are not shown here.

Figure 5 shows variation of the measured gain of the developed antennas across the ultra wideband. It is clear from the results in Figure 5 that the antennas have a low gain, which coincides with the expected behavior of the omnidirectional antennas. The gain of the antenna without a slot increases with frequency from 0.3 dB at 3 GHz to 3.1 dB at 9 GHz, and then it decreases a little and becomes 2.7 dB at 11 GHz. Concerning gain of the developed antenna with tuning slot, the measured results, presented in Figure 5, show that the general behavior of the gain is similar to that of the antenna without tuning slot except for the rejected subband. Across the band 4.9–5.9 GHz, the gain can be as low as \(-2.3 \text{ dB}\).
when a tuning slot is used, compared with 2.4 dB for the antenna without a slot. This proves the high capability of the added slot to reject the undesired subband.

The last investigation concerns variations of the radiation efficiency for the designed antennas with and without the tuning slot. Results of the calculations using the software HFSS indicated that the investigated antennas feature a good efficiency, being greater than 90% across the desired band.

4. CONCLUSION

In this article, a simple design method for a planar ultra wideband antenna has been presented. The presented antenna covers the frequency assigned by FCC for ultra wideband applications and rejects the band assigned for other applications. The proposed radiating element and the ground plane are of an elliptical shape with a coplanar waveguide feeder. The results presented in this article have shown that the designed antenna has a 10 dB return loss bandwidth from 2.8 GHz to more than 11 GHz. The results have also revealed that the undesired subband used by other applications can be excluded from the response of the antenna by making a tuning slot, with length, width, and position calculated according to the design method, in the radiator of the antenna. The designed antenna has shown omnidirectional radiation pattern and more than 90% radiation efficiency in the pass-band.

Figure 4  Radiation pattern of the manufactured antenna without a slot at different frequencies. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

Figure 5  Variation of the gain with frequency for the manufactured antennas
A NOVEL DUAL-BAND BRANCH LINE COUPLER BASED ON STRIP-SHAPED COMPLEMENTARY SPLIT RING RESONATORS

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ABSTRACT: In this article, a novel dual-band branch line coupler is presented. The dual-band operation is realized by the LH-RH transmission line composed of strip-shaped complementary split ring resonators (CSRRs) and series gap, which can produce LH band around 900 MHz and RH band around 1800 MHz. The performances of the designed device at two operating frequencies are demonstrated by simulated and measured results, which are in good agreement. © 2007 Wiley Periodicals, Inc. Microwave Opt Technol Lett 49: 2859–2862, 2007; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.22873

Key words: strip-shaped; complementary split ring resonators; left-handed materials; dual-band; branch line coupler

1. INTRODUCTION

In the recent years a new type of metamaterials called left-handed materials (LHM) has attracted many attentions. These materials with effective negative permittivity and permeability simultaneously are first realized [1] by arranging arrays of thin metallic continuous wires and split ring resonators (SRRs) [2]. LHM supports propagation of backward waves whose phase velocity is opposite to the direction of energy flow. From duality, negative permittivity medium can be generated by complementary split ring resonators (CSRRs) [3]. Marques and co-workers first introduced these two resonators into planar structures and a LH pass-band can be found, while CSRRs combined with a series gap are fabricated in microstrip technology and SRRs combined with a shunt strip are fabricated in coplanar waveguide technology [4–6]. By improving the series capacitance, the LH transmission band and RH transmission band will overlap and bandwidth is increased [7, 8].

There are two another completely independent approaches based on the theory of transmission lines to make LHM. One is to use the interdigital capacitors and short-circuit stub inductors [9], and the other is to use surface-mount-technology (SMT) lumped elements (LEs) [10]. By using the composite LH and RH characteristics implemented by the LEs, dual-band branch-line coupler and rat-race coupler are realized [11].

In this article, a novel method is proposed to design dual-band branch-line coupler, in which strip-shaped CSRRs combined with series capacitance are used to produce LH and RH transmission band. Two operating frequencies are chosen as 900 MHz and 1800 MHz which are used in global systems for mobile communications (GSM). Simulated and measured results are in good agreement. Because no lumped elements and grounded via are needed, this device is easier to be fabricated compared with the dual-band branch line coupler adopting lumped elements.

2. LH AND RH BAND

The layout of strip-shaped CSRRs combined with series capacitance implemented by a series gap and its lumped element equivalent T-circuit model are shown in Figure 1. As proposed in [8], under the assumption that the electrical size is small compared with the wavelength and considering the part as the basic cell of periodical structure, the phase shift factor can be obtained:

\[ \cos \phi = 1 + \frac{C(1 - \omega^2LC_g)(1 - \omega^2L_cC_g)}{2C_g[1 - \omega^2L_cC_g]} \]  

(1)

Eq. (1) indicates that the left-hand pass-band occurs in the frequency region:

\[ f_L = \frac{1}{2\pi} \sqrt{\frac{C + 4C_g}{L(4C_g(C_c + C) + C(C_c + L))}} \]  

(2)

\[ f_H = \frac{1}{2\pi} \sqrt{\frac{C}{L_cC_g}} \]  

(3)

Right-hand pass-band can be found above the frequency:

\[ f_H = \frac{1}{2\pi} \sqrt{\frac{C}{L_cC_g}} \]  

(4)

By adjusting the series capacitance \( C_c \), right-hand pass-band can be moved, and especially as \( C_c \) is increased to a certain value a balanced CRLH transmission with wide pass-band can be produced, which give us the possibility that 900/1800 MHz dual-band

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Design of a Compact Ultra Wideband Antenna with Interdigital Resonator for Subband Rejection

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Abstract — A complete design method for a compact ultra wideband planar antenna with subband rejection capability is presented. An interdigital resonator is incorporated in the microstrip feeder of the antenna to act as a bandstop filter, hence enabling the rejection of any undesired band within the passband of the antenna. Two samples of the proposed antenna were designed and manufactured. One of the developed antennas does not contain a resonator, whereas the other contains an interdigital resonator. The designed antennas feature a compact size of 25 mm × 25 mm. Results of the simulation and measurement show that the designed antennas have a bandwidth from 3 GHz to more than 11 GHz. The results also show that the use of the resonator in the microstrip feeder of the antennas efficiently rejects any undesired subband, such as the 4.9-5.9 GHz band assigned for IEEE802.11a and HIPERLAN/2. The gain of the antennas with resonator is about 2.5 dB at the passband, while it is less than -10 dB at the rejected subband.

Index Terms— Ultra wideband antenna, interdigital resonator, subband rejection.

I. INTRODUCTION

Ultra wideband (UWB) systems have received a considerable amount of interest with respect to communication and medical applications after the Federal Communication Commission in the USA permitted the marketing and operation of UWB within the range 3.1 GHz to 10.6 GHz. An antenna, which can efficiently radiate and receive UWB signals, is essential for the successful operation of these systems. In addition to the requirement of the 3.1-10.6 GHz bandwidth, the UWB antennas should have the capability to reject any interference with existing wireless networking technologies such as the subband 4.9-5.9 GHz for IEEE 802.11a in the USA, and HIPERLAN/2 in Europe. Therefore, UWB antennas with notched characteristics at certain bands are desired. Some designs have appeared recently for UWB antennas with subband rejection capability [1-3]. The rejection capabilities of the antennas presented in those papers were achieved by creating a slot in the radiating element of the antennas. Although the slot is usually designed to filter out the undesired band while keeping the desired band intact, the measured results, for example in [3], show that there is some negative impact on the return loss of the antenna at the high end of the passband.

This paper describes a different design method for a UWB antenna with a subband rejection capability. An interdigital resonator is incorporated in the microstrip feeder of the antenna to filter out any undesired subband with negligible effect on the passband of the antenna. The method presented in this paper, is used to design a UWB antenna which rejects the 4.9-5.9 GHz band, which is assigned for the IEEE802.11a and HIPERLAN/2. The results of measurements for two manufactured antennas (one without a resonator and the other with an interdigital resonator) are presented in order to prove the validity of the proposed method.
II. DESIGN

The configuration of the proposed UWB antenna is illustrated in Fig.1. The radiating structure is formed from a circular strip. The antenna is fed using a microstrip line of width $w_m$, which can be calculated using the microstrip design equations [4]. The ground plane, which is located at the other side of the substrate, is in the shape of half a circle. To reject a certain subband within the passband of the antenna, an interdigital resonator is incorporated in the microstrip feeder of the antenna in the manner shown in Fig.1a.

![Fig.1. (a) Configuration of the proposed antenna and (b) details of the interdigital resonator.](image)

Depend on the lowest frequency of operation ($f_l = 3.1$ GHz) and the dielectric constant of the substrate ($\varepsilon_r$), width ($w$) and length ($l$) of the antenna structure are assumed to be equal to half of the effective wavelength calculated at $f_l$:

$$w = l = \frac{c}{2f_l \sqrt{(\varepsilon_r + 1)/2}} \quad (1)$$

The ground plane of the antenna is in the shape of half a circle with diameter equal to $w$. The radiating structure, which is located at a distance $g$ from the end of the ground plane, is in the form of a circular strip. The outer diameter of the strip is equal to $w$, whereas the inner diameter is chosen such that width of the radiator is equal to $w_r$. A parametric analysis on the effect of $w_r$ on the performance of the antenna indicated that it has a negligible effect on the lower and upper frequency response of the antenna’s passband, while increasing $w_r$ improves the return loss of the antenna at the centre of the passband. This means that if the antenna is designed to have an UWB performance without rejecting any subband then it is better to increase $w_r$ to as much as possible, depending of course on size of the antenna structure. Because the antenna designed in this paper is aimed to have a rejection capability at a certain subband near the centre of its passband, then $w_r$ is to be chosen such that the standing wave ratio (SWR) at that subband is around 2 (i.e. -10 dB return loss). Hence, the overall effect of $w_r$ and the interdigital resonator, which is to be explained thereafter, results in an optimum rejection of the undesired subband.

Concerning the gap value $g$ between the radiator and the ground plane, a parametric analysis on the best value for $g$ indicated that it should be comparable to thickness of the substrate $h$ in order to get the widest bandwidth. In the design procedure of this paper, $g$ is assumed to be equal to $h$.

The above design procedure results in an antenna which covers the whole UWB range from 3.1 GHz to 10.6 GHz. If it is required to reject a certain subband within that range, an interdigital resonator can be incorporated in the microstrip feeder of the antenna in the manner shown in Fig.1a. This resonator consists of an interdigital capacitor in parallel with a pair of lumped-element inductors. Configuration of the resonator is shown in Fig.1b.

The interdigital capacitor was originally proposed by Alley [5] for use in lumped-element microwave integrated circuits. Letting the finger width $w_c$ equal the space $s$ to achieve maximum capacitance density, and assuming that the substrate thickness $h$ is much larger than the finger width, a closed-form expression for estimation of capacitance of the interdigital capacitor is given by [5]:

$$C = 3.937 \times 10^{-14} l_r (\varepsilon_r + 1)[0.1 \log(n-3) + 0.252] \quad (2)$$
where \( l_c \) is length of the interdigital capacitor in (mm) and \( n \) is number of the fingers in the capacitor.

Concerning the outer strips of the resonator shown in Fig.1b, they act as parallel inductors. An approximate design equation can be used to find value of the inductance of each strip depending on its width \( w_i \) and length \( l_r \) \[6-7]\:

\[
L = 2 \times 10^{-10} l_r \left[ \ln \left( \frac{l_r}{w_i} \right) + 1.193 + \frac{0.2235 w_i}{l_r} \right] \times k \quad (3)
\]

\[
k = 0.57 - 0.145 \ln \left( \frac{w_i}{h} \right) \quad (4)
\]

The resonator shown in Fig.1a has two inductors and an interdigital capacitor connected in parallel. Therefore, the resonance frequency is equal to:

\[
f_r = \frac{1}{2\pi \sqrt{LC/2}} \quad (5)
\]

At resonance, the interdigital resonator used in this paper behaves like an open circuit because of the parallel resonance. Hence, the frequency given in (5) can be assumed as the centre of the rejected subband. The design parameters for the resonator can be found using (2)-(5) for a certain rejected subband. In this paper, the centre of the rejected subband is considered to be \((4.9+5.9)/2=5.4 \) GHz in order to tune out any interference caused by the subbands used by IEEE802.11a and HIPERLAN/2.

### III. RESULTS

The proposed UWB antenna was designed and manufactured using Rogers 4003C substrate (\( \varepsilon_r=3.38, \ \tan\delta=0.00023, \) and thickness=0.508 mm). Values of the design parameters for the antenna and the incorporated resonator were first calculated using the proposed design procedure and then optimized using the software Ansoft HFSSv10. The optimized values are; \( w=25 \) mm, \( l=25.5 \) mm, \( g=0.5 \) mm, \( w_r=6 \) mm, \( w_m=1.17 \) mm, \( l_r=5.5 \) mm, \( w_c=0.13 \) mm, \( s=0.13 \) mm, and \( w_y=0.13 \) mm.

Two samples of the proposed antenna were manufactured, one without and the other with the interdigital resonator. Characteristics of the developed antennas were tested via simulations using the software HFSSv10 and via measurements using a vector network analyser in an anechoic chamber. Fig.2 shows variation of the SWR with frequency for the developed antennas. The simulated and measured characteristics of the antenna without a resonator reveal UWB behaviour with bandwidth from 3 GHz to more than 11 GHz assuming SWR=2 (or 10 dB return loss) as a reference. It is also clear from Fig.2 that, for the antennas with a resonator, the undesired subband was tuned out, whereas the wideband behaviour of the antenna was maintained. The simulated and measured results are in good agreement.

![Fig.2. Variation of the SWR with frequency.](image)

From the UWB applications point of view, the antenna is usually required to have an omnidirectional radiation. Concerning the designed antenna, this requirement is fulfilled over the whole bandwidth as shown in Fig.3 for the antenna without a resonator. The radiation patterns of the antenna with a resonator are similar to the results shown in Fig.3 and therefore, they are not shown here.

![Fig.3. Three dimensional radiation pattern of the antenna at different frequencies.](image)
expected behavior of the omnidirectional antennas. The gain of the antenna without a resonator increases with frequency from around 0 dB at 3.1 GHz to 4.3 dB at 9 GHz, and then it decreases a little and becomes 3 dB at 11 GHz. Concerning gain of the developed antenna with the interdigital resonator, the measured results, presented in Fig.4, show that the general behavior of the gain is similar to that of the antenna without a resonator except for the rejected subband. Across the band 4.9 GHz to 5.9 GHz, the gain can be as low as -10 dB, when a resonator is used, compared with 2.5 dB for the antenna without a resonator. This proves the high capability of the proposed resonator to reject the undesired subband.

Fig.4. Variation of the measured gain with frequency for the manufactured antennas.

If the antenna is to be used in UWB pulse transmission/reception then the group delay of the antenna is of utmost importance as it indicates the level of distortion in the received pulse. Concerning the designed antenna, the simulation results indicated that the antenna has an almost flat group delay across its passband with a maximum deviation of less than 1 ns.

The last investigation concerns variation of the radiation efficiency for the designed antennas with and without the resonator. Results of the calculations using the software HFSS indicated that the proposed antennas feature a good efficiency, being greater than 90% across the desired band.

IV. CONCLUSION

The design of a compact ultra wideband planar antenna with subband rejection capability has been presented. An interdigital resonator is incorporated in the microstrip feeder of the antenna to act as a bandstop filter, hence enabling the rejection of any undesired band within the passband of the antenna. Two samples of the proposed antenna were designed and manufactured. One of the developed antennas does not contain a resonator, whereas the other contains an interdigital resonator. The designed antennas feature a compact size of 25 mm × 25 mm. Results of the simulation and measurement show that the designed antennas have a bandwidth from 3 GHz to more than 11 GHz. The results also show that the use of the resonator in the microstrip feeder of the antennas efficiently rejects any undesired subband, such as the 4.9-5.9 GHz band assigned for IEEE802.11a and HIPERLAN/2. The gain of the antennas with the resonator is about 2.5 dB at the passband, while it is less than -10 dB at the rejected subband.

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Research Article

Design of a CPW-Fed Band-Notched UWB Antenna Using a Feeder-Embedded Slotline Resonator

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A complete design method for a compact uniplanar ultra-wideband antenna with subband rejection capability is presented. A slotline resonator is incorporated in the coplanar waveguide feeder of the antenna to act as a bandstop filter, hence enabling the rejection of any undesired band within the passband of the antenna. Two samples of the proposed antenna were designed and manufactured. One of the developed antennas does not contain a resonator, whereas the other contains a slotline resonator. The designed antennas feature a compact size of 27 mm × 27 mm. Results of the simulation and measurement show that the designed antennas have a bandwidth from 3 GHz to more than 11 GHz. The results also reveal that the use of the resonator in the feeder of the antenna efficiently rejects any undesired subband, such as the 4.9–5.9 GHz band assigned for IEEE802.11a and HIPERLAN/2. The gain of the antennas with the resonator is about 2.2 dBi at the passband, while it is less than −8 dBi at the rejected subband.

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1. INTRODUCTION

Ultra-wideband (UWB) is a short range communications technology. It has gained a lot of popularity among researchers and the wireless industry after the FCC permitted the marketing and operation of UWB within the range 3.1 GHz to 10.6 GHz [1]. The global interest in the UWB technology is increasing very fast due to the capability of this license exempt wide bandwidth system to yield low cost, low energy, short range, and extremely high capacity wireless communication links.

In addition to the requirement of the 3.1–10.6 GHz bandwidth, the UWB antennas should also have the capability to reject any interference with existing wireless networking technologies such as the subband 4.9–5.9 GHz for IEEE 802.11a in the USA, and HIPERLAN/2 in Europe. Therefore, UWB antennas with notched characteristics at certain bands are desired. The ability to provide this function in the antenna can significantly relax the requirements imposed upon the filtering electronics within the wireless communication system.

One of the first attempts to design a band-notched ultra-wideband antenna was presented in [2]. The design of a UWB antenna with a good impedance matching over the desired band was achieved using a genetic algorithm optimization code. A manual trial-and-error approach was then utilized to modify the design towards a band-notched antenna. In a later stage, the authors of [2] extend their optimization technique to improve the radiation pattern of their band-notched design [3].

A systematic method for designing frequency notched ultra-wideband antennas was presented in [4]. By deliberately introducing a narrow band resonant structure, an antenna may be made capable of rejecting particular frequencies. This technique is useful for creating UWB antennas with narrow frequency notches, or for creating multiband antennas.

Reviewing the literature shows that there are many different methods used to achieve the band-notched function. The conventional methods are cutting a slot of different shapes on the radiating patch [5–7], inserting a slit on the patch [8], embedding a quarter-wavelength tuning stub within a large slot on the patch [9], putting parasitic elements near the radiator as filters to reject the limited band [10], or introducing a parasitic open-circuit element, rather than modifying the structure of the antenna’s tuning stub [11].
This paper describes a different design method for a UWB antenna with a subband rejection capability. Instead of modifying the radiator as with the proposed methods in the literature, a slotline resonator is incorporated in the coplanar waveguide feeder of the antenna to filter out any undesired subband with negligible effect on the passband of the antenna. The method presented in this paper is used to design a uniplanar UWB antenna which rejects the 4.9–5.9 GHz band, which is assigned for the IEEE802.11a and HIPERLAN/2. The results of simulations and measurements for two manufactured antennas (one without a resonator and the other with a slotline resonator) are presented in order to prove the validity of the proposed method.

2. DESIGN

The configuration of the proposed UWB antenna with the capability of rejecting frequencies over a certain subband is illustrated in Figure 1. The radiating structure is formed by the connection of half an ellipse with a rectangular patch. The antenna is assumed to be fed using a coplanar waveguide (CPW) to enhance its broadband characteristics. The ground plane, which is located around the CPW feeder, is in the shape of half an ellipse.

The design procedure starts by finding dimension of the antenna feeder to give 50 Ω characteristic impedance (Z₀). This can be achieved using the following equations [12]:

\[
Z_0 = \frac{30\pi K'(k)}{\sqrt{\varepsilon_r}} K(k),
\]

\[
k = \frac{s}{s + 2w_c},
\]

\[
\varepsilon_r = \varepsilon_r' + \frac{1}{2} [A + B],
\]

\[
A = \tanh \left[ 1.785 \log (h/w_c) + 1.75 \right],
\]

\[
B = \frac{k w_c}{h} \left( 0.04 - 0.7k + 0.01(1 - 0.1\varepsilon_r')(0.25 + k) \right),
\]

where \( K(k) \) is the first kind elliptical integral and \( K'(k) = K(\sqrt{1 - k^2}) \), \( s \) is the central conductor width, whereas \( w_c \) is the slot width of the CPW, \( h \) is the substrate thickness, and \( \varepsilon_r \) is the dielectric constant of the substrate.

Depending on the lowest frequency of operation (\( f_l = 3.1 \) GHz), thickness of the substrate and its dielectric constant, width (\( w \)), and length (\( l \)) of the antenna structure are calculated as

\[
w = l = \frac{c}{2f_l \sqrt{\varepsilon_r}}.
\]

The ground plane of the antenna, which is located around the CPW, is half of an ellipse with major diameter equal to (\( w \)). The secondary diameter of the ground plane can be chosen to be around (0.5 \( w \)).

The radiating structure, which is located at a distance \( g \) from the end of the ground plane, consists of two parts. The first part, which is connected directly with the feeder, is half an ellipse with the same dimensions as for the ground plane. The second part is a rectangular patch which extends from the end of the half ellipse till the end of the substrate. Parametric analysis on the best value for \( g \) indicates that it should be less than thickness of the substrate \( h \) in order to get the widest bandwidth. In the design procedure of this paper, \( g \) is assumed to be equal to \( (h/3) \).

It is worthwhile to mention that (2) gives an accurate estimation of the required dimension of the antenna for a low value of the dielectric constant of the substrate, that is, less than 4. If it is required to find a rough estimation of the required dimension for any value of the dielectric constant, the following formula can be used:

\[
w = l = \frac{c}{2f_l(\sqrt{\varepsilon_r + 1})/2}.
\]

The explained design procedure (1)–(3) results in an antenna which covers the whole UWB range from 3.1 GHz to 10.6 GHz. If it is required to reject a certain subband within...
that range, a slotline resonator can be incorporated in the feeder of the antenna in the manner shown in Figure 1(a). Configuration of the resonator is shown in Figure 1(b). Length of the slotline resonator is assumed to be equal to a quarter of the effective wavelength calculated at the center of the rejected band.

To understand effect of the slotline used in this paper, a transmission line model of the coplanar waveguide and the slotline resonator is shown in Figure 1(c). The resonator is equivalent to a short-circuited end series stub. At the center of the rejected band, the stub’s length is equal to a quarter wavelength which means that the stub appears effectively as an open circuit at its point of connection with the main transmission line, which is the coplanar waveguide for the design presented in this paper. This comes from the fact that the input impedance of a quarter-wavelength stub is equal to $Z_o^2/Z_o$ [13], where $Z_o$ is the characteristic impedance of the stub and $Z_t$ is the terminal impedance of the stub, which is equal to zero in the stub considered in this paper. At the frequencies, which are far away from the resonance frequency of the stub, the resonator has negligible effect on the performance and the feeder and hence, the antenna performs as if there is no resonator.

It is to be noted that in order to use the space available for the stub efficiently and to make the antenna compact in size, the stub is folded in the manner shown in Figure 1(b).

### 3. RESULTS

The proposed UWB antenna was designed using GML1032 substrate with a dielectric constant equal to 3.2, tangent loss $\tan\delta = 0.004$, and thickness of 1.52 mm. Values of the design parameters for the antenna and the incorporated resonator were first calculated using the proposed design procedure and then optimized using the software Ansoft HFSSv10. The optimized values are: $w = 17$ mm, $s = 2.9$ mm, $w_s = 0.25$ mm, $g = 0.5$ mm, $l_s = 11$ mm, $w_t = 0.3$ mm, and $s_t = 0.1$ mm.

Two samples of the proposed antenna were manufactured, one without and the other with the slotline resonator. Characteristics of the developed antennas were tested via simulations using the software HFSSv10 and via measurements using a vector network analyser in an anechoic chamber. Figure 2 shows variation of the SWR with frequency for the developed antennas. The simulated and measured characteristics of the antenna without a resonator reveal UWB behavior with bandwidth from 3 GHz to more than 11 GHz assuming SWR = 2 (or 10 dB return loss) as a reference. It is also clear from Figure 2 that, for the antennas with a resonator, the undesired subband was tuned out, whereas the wideband behavior of the antenna was maintained. The simulated and measured results are in good agreement.

It is worthwhile to mention that the slot width of the resonator $s_t$ can be used to adjust width of the rejected subband. A parametric analysis using the software HFSS shows that increasing $s_t$ increases width of the rejected band and its level of rejection, that is, a higher VSWR at the rejected band with a larger slot width. Moreover, the parametric analysis shows that width of the part of the feeder bounded by the resonator, that is, $w_t$ in Figure 1(b), has a significant effect on the level of rejection. Optimization techniques can be utilized to achieve a certain compromise between the required width of the rejected band and the level of rejection.

From the UWB applications point of view, the antenna is usually required to have an omnidirectional radiation. Concerning the designed antenna, this requirement is fulfilled over the whole bandwidth as shown in Figure 3 for the antenna without a resonator. The radiation patterns of the antenna with a resonator are similar to the results shown in Figure 3 and therefore, they are not shown here.

Figure 4 shows variation of the measured gain of the developed antennas across the ultra-wideband. It is clear from the results in Figure 4 that the antennas have a low gain, which agrees with the expected behavior of the omnidirectional antennas. The gain of the antenna without a resonator increases with frequency from around 0 dBi at 3.1 GHz to 3 dBi at the range 9–11 GHz. Concerning gain of the developed antenna with the slotline resonator, the measured results, presented in Figure 4, show that the general behavior of the gain is similar to that of the antenna without a resonator except for the rejected subband. Across the band 4.9 GHz to 5.9 GHz, the gain can be as low as -8 dBi, when a resonator is used, compared with 2.2 dBi for the antenna without a resonator. This proves the high capability of the proposed resonator to reject the undesired subband.

The last investigation concerns variation of the radiation efficiency for the designed antennas with and without the resonator. Results of the calculations using the software HFSS indicated that the proposed antennas feature a good efficiency, being greater than 92% across the desired band.
4. CONCLUSION

This paper has presented a complete design method for a compact ultra-wideband antenna with band-notched characteristics. The subband rejection is accomplished by incorporating a slotline resonator in the coplanar waveguide feeder of the antenna. To validate the proposed method, two samples of the proposed antenna were designed and manufactured: one does not contain a resonator, whereas the other contains a slotline resonator. Results of the simulation and measurement have shown that the designed antennas have an ultra-wideband performance with bandwidth from 3 GHz to more than 11 GHz. The results have also shown effectiveness of the resonator in rejecting any undesired band, such as the 4.9-5.9 GHz band, where the gain of the antenna with the resonator was reduced by more than 10 dB at the rejected band.

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ULTRA WIDEBAND PLANAR ANTENNA WITH SPURLINE FOR SUBBAND REJECTION

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ABSTRACT: The design of a compact ultra wideband planar antenna with subband rejection capability is presented. A spurline is incorporated in the microstrip feeder of the antenna to act as a bandstop filter, hence enabling the rejection of any undesired band within the passband of the antenna. Results of a parametric investigation are presented to show effect of the spurline parameters on width of the rejected subband. The investigation shows that increasing the slot and spurline widths increases width of the rejected subband. Three samples of the proposed antenna were designed and manufactured. One of the developed antennas does not contain spurline, whereas the other two contain spurlines with different dimensions. The designed antennas feature a compact size of 25 mm × 25 mm. Results of the simulation and measurement show that the designed antennas have a bandwidth from 3 GHz to more than 11 GHz. The results also show that the use of the spurlines in the microstrip feeder of the antennas efficiently rejects any undesired subband, such as the 4.9–5.9 GHz band assigned for IEEE802.11a and HIPERLAN/2. The gain of the antennas with spurline is about 2.5 dB at the passband, while it is less than −15 dB at the rejected subband. © 2008 Wiley Periodicals, Inc. Microwave Opt Technol Lett 50: 725–728, 2008; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.23205

Key words: planar antenna; subband rejection; ultra wideband antenna

1. INTRODUCTION

Design of ultra wideband (UWB) antennas is receiving an increased attention from the researchers and the wireless industry after the Federal Communication Commission in the USA permitted the marketing and operation of UWB within the range 3.1 GHz to 10.6 GHz. In addition to the requirement of the 3.1–10.6 GHz bandwidth, the UWB antennas should have the capability to reject any interference with existing wireless networking technologies such as the subband 4.9–5.9 GHz for IEEE 802.11a in the USA, and HIPERLAN/2 in Europe [1]. Therefore, UWB antennas with notched characteristics at certain bands are desired. Some designs have appeared recently for UWB antennas with subband rejection capability [2-4]. The rejection capabilities of the antennas presented in those articles were achieved by creating a slot in the radiating element of the antennas. Although the slot is usually designed to filter out the undesired band while keeping the desired band intact, the measured results show that it has a negative impact on the return loss of the antenna at part of the passband.

This article describes a different design method for a UWB antenna with a subband rejection capability. A spurline is incorporated in the microstrip feeder of the antenna to filter out any undesired subband with negligible effect on the passband of the...
The above design procedure results in an antenna which covers the whole UWB range from 3.1 GHz to 10.6 GHz. If it is required to reject a certain subband within that range, a spurline can be incorporated in the microstrip feeder in a manner shown in Figure 1(a). As the spurline has a high rejection capability across its stopband and a very low insertion loss across its passband [6], then incorporating it into the antenna’s feeder is expected to have negligible effect on the performance across the passband, while it filters out the undesired subband. The spurline consists of a pair of coupled microstrip lines with one line open ended; both lines are connected together at the other end. Configuration of the spurline is shown in Figure 1(b).

The spurline is described by three parameters: slot length \( l_s \), slot width \( s \), and spurline width \( w_s \). See Figure 1(b). Length of the slot defines center of the rejected subband and it is equal to quarter of the effective wavelength calculated at that frequency. The performance of the spurline can be analyzed assuming a quasi TEM propagation modes and using the even- and odd-mode theory of the coupled lines [7]. According to that theory, the slot width \( s \) and the spurline width \( w_s \) define value of the even- and odd-mode impedances which, in turn, define the frequency response of the spurline.

A parametric analysis was used to show effect of \( s \) and \( w_s \) on width of the rejected subband. Rogers4003C (dielectric constant = 3.38, tangent loss \( \tan \delta = 0.00023 \), and thickness = 0.508 mm) was assumed as a substrate for the purpose of the analysis. Figure 2 shows the results of the analysis. It is to be noted that the fractional bandwidth of the rejected subband with respect to its centre was calculated assuming the 3 dB insertion loss as a reference. The slot and spurline widths have a direct effect on value of the coupling between the two coupled lines of the spurline. Therefore, they define width of the rejected subband. The results shown in Figure 2 indicate that the smaller are the widths \( s \) and \( w_s \), the narrower is the rejected subband. Increasing width of the gap or the spurline increases width of the rejected subband till a certain value after which, increasing the widths has no effect any more on width of the rejected subband. The gap between the input microstrip line and the open-circuited line of the coupler has a negligible effect on the performance of the filter. Therefore, it is considered to be approximately equal to the slot width.

Figure 2: Effect of the spurline’s dimensions on the fractional bandwidth of the rejected subband. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
3. RESULTS

The proposed UWB antenna was designed and manufactured using Rogers4003C substrate ($\varepsilon_r = 3.38$, $\delta = 0.00023$, and thickness = 0.508 mm). Values of the design parameters for the antenna were first calculated using the proposed design procedure and then optimized using the software Ansoft HFSSv10. The optimized values are as follows: $w = 25$ mm, $l = 25.5$ mm, $g = 0.5$ mm, $w_s = 6$ mm, and $w_m = 1.18$ mm.

Concerning the spurline, assume that it is required to design the antenna to reject the subband 4.9 GHz to 5.9 GHz from its response. The center frequency of the undesired subband $f_c$ is equal to $(4.9 + 5.9)/2 = 5.4$ GHz. Therefore, length of the spurline is equal to 8.7 mm, which is quarter of the effective wavelength at the center of the rejected subband. The fractional bandwidth of the rejected subband is equal to $(5.9 – 4.9)/5.4 = 18.5\%$. From Figure 2, it is clear that there are several combinations of $s$ and $w_s$ values to achieve the required fractional bandwidth of the rejected sub-band. One of them is as follows: $s = w_s = 0.2$ mm. If it is required to reject a narrower band then lower values for $s$ and $w_s$ can be used. As an example for this case, another antenna with $s = w_s = 0.1$ mm was also manufactured and tested.

Three samples of the proposed antenna were manufactured, one without a spurline and the other two with spurline1 (with $s = w_s = 0.2$ mm) and spurline2 ($s = w_s = 0.1$ mm). Characteristics of the developed antennas were tested via simulations using the software HFSSv10 and via measurements using a vector network analyzer in an anechoic chamber. Figure 3 shows variation of the SWR with frequency for the developed antennas. The simulated and measured characteristics of the antenna without a spurline reveal UWB behavior with bandwidth from 3 GHz to more than 11 GHz assuming SWR = 2 (or 10 dB return loss) as a reference. It is also clear from Figure 3 that, for the antennas with a spurline, the undesired subband was tuned out, whereas the wideband behavior of the antenna was maintained. Antenna with spurline1 shows a wider filtered subband compared with the antenna with spurline2 which has lower values for the slot and spurline widths. The simulated and measured results are in good agreement. The width of the rejected subbands in the simulated and measured results of Figure 3 agrees well with the results of the parametric analysis shown in Figure 2.

From the UWB applications point of view, the antenna is usually required to have an omnidirectional radiation. Concerning the designed antenna, this requirement is fulfilled over the whole bandwidth as shown in Figure 4 for the antenna without a spurline. The radiation patterns of the antennas with a spurline are similar to the results shown in Figure 4 and therefore, they are not shown here.

Figure 5 shows variation of the measured gain of the developed antennas across the ultra wideband. It is clear from the results in Figure 5 that the antennas have a low gain, which coincides with the expected behavior of the omnidirectional antennas. The gain of
the antenna without a spurline increases with frequency from 0 dB at 3.1 GHz to 4.3 dB at 9 GHz, and then it decreases a little and becomes 3 dB at 11 GHz. Concerning gain of the developed antennas with spurline, the measured results, presented in Figure 5, show that the general behavior of the gain is similar to that of the antenna without tuning slot except for the rejected subband. Across the band 4.9 GHz to 5.9 GHz, the gain can be as low as −15 dB, when a spurline is used, compared with 2.5 dB for the antenna without a spurline. This proves the high capability of the spurline to reject the undesired subband.

The last investigation concerns variation of the radiation efficiency for the designed antennas with and without the spurline. Results of the calculations using the software HFSS indicated that the investigated antennas feature a good efficiency, being greater than 90% across the desired band.

4. CONCLUSION

The design of a compact ultra-wideband planar antenna with subband rejection capability has been presented. A spurline is incorporated in the microstrip feeder of the antenna to act as a bandstop filter, hence enabling the rejection of any undesired band within the passband of the antenna. The results of a parametric investigation have been presented to show effect of the spurline parameters on width of the rejected subband. The investigation has revealed that increasing the slot and spurline widths increases width of the rejected subband.

Three samples of the proposed antenna were designed and manufactured. One of the developed antennas does not contain spurline, whereas the other two contain spurlines with different dimensions. The designed antennas feature a compact size of 25 mm × 25 mm. Results of the measurement have shown that the designed antennas have a bandwidth from 3 GHz to more than 11 GHz. The results have also shown that the use of the spurlines in the microstrip feeder of the antennas efficiently rejects any undesired subband, such as the 4.9–5.9 GHz band assigned for IEEE802.11a and HIPERLAN/2. The gain of the antennas with spurline is about 2.5 dB at the passband, while it is less than −15 dB at the rejected subband.

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tially relaxed by introducing the band rejection function within the wireless transceiver. This calls for allocating a suitable area for suppressing bands. The time domain transmission test between two identical antennas without the parasitic strips shows an almost distortionless pulse performance.

Index Terms—Band-notched antenna, planar antenna, ultra wideband antenna.

I. INTRODUCTION

Ultrawideband (UWB) technology has gained a lot of popularity among researchers and the wireless industry after the FCC permitted its marketing within the frequency band of 3.1 GHz to 10.6 GHz [1]. The attractiveness of UWB is in its capability of offering high capacity short-range wireless communication links using low-cost low-energy transceivers. To establish the communication between two nodes, these transceivers require UWB antennas, preferably of small size and low manufacturing cost. Planar monopole antennas of various shapes and feeding structures (coaxial, microstrip and coplanar waveguide type) have been found as good candidates to fulfill this requirement.

Because of the existence of other wireless standards, such as IEEE 802.11a or HIPERLAN/2 operating in the 4.9–5.9 GHz, an additional requirement for UWB antennas is to reject some bands within the ultra wide passband. In these cases, UWB antennas with notched characteristics at certain bands are desired. The function of rejecting certain frequencies can be accomplished within the wireless transceiver by employing a band rejection filter. This calls for allocating a suitable area within the transceiver for such a device. This requirement can be significantly relaxed by introducing the band rejection function within the UWB antenna structure.

The literature review shows a number of different methods that can be used to achieve the band-notch function. One of the first attempts to design a band-notched ultra wideband antenna was presented in [2]. The design of a UWB antenna with a good impedance matching over the desired band was achieved using a genetic algorithm optimization code. A manual trial-and-error approach was then applied to modify the design towards a band-notched antenna. In a later stage, the authors of [2] extended their optimization technique to improve the radiation pattern of the presented band-notched design [3].

The other band rejection methods reported in the literature that can be applied to a UWB antenna include cutting a slot of different shapes on the radiating patch [4]–[7], inserting a slit on the patch [8], embedding a quarter-wavelength tuning stub within a large slot on the patch [9], putting parasitic patches near the radiator as filters to reject the limited band [10] or introducing a parasitic open-circuit element [11]. In addition, in [12] the split ring resonator was utilized within the radiator to achieve the required band rejection. In [13], the combined effect of ground plane deformation and slots embedded within the radiator was employed to build an UWB antenna with band-notch characteristics. However, the use of embedded slots within the radiator had the negative impact on the gain of the antenna. The gain across a significant part of the passband was below −2 dBi. Also, the use of the deformed ground plane made it difficult to integrate the antenna with the RF frontend.

This communication describes a new method for the design of a UWB planar antenna with band-notch characteristics. In this method, parasitic elements in the form of printed strips placed in the radiating aperture of the planar antenna at the top and bottom layer are employed to suppress the radiation at certain frequencies within an ultra wide frequency band. The parasitic elements have dimensions which are chosen according to a certain formula. They can be used to reject a single narrow band, a wide band, or three narrow bands, while the normal performance of the antenna is maintained at the remaining passband. The effectiveness of the proposed method compares favorably with the other methods aiming at the band rejection within UWB. The voltage standing wave ratio (VSWR) at the rejected bands is more than 30, while the antenna gain is reduced by more than 10 dB.

II. DESIGN

The configuration of the investigated planar UWB antenna with the capability of rejecting frequencies over a certain band is illustrated in Fig. 1(a). The radiating structure is in the form of half circle. The ground plane located at the reverse side of the substrate is also in the shape of half circle. The antenna is fed using a microstrip line whose width is calculated using the well-known microstrip line design equations [14].

Depending on the lowest frequency of operation \( f_{\text{lo}} = 3.1 \, \text{GHz} \), thickness of the substrate and its dielectric constant \( \varepsilon_r \), width \( w \) and length \( l \) of the antenna structure are calculated as:

\[
 w = l = \frac{c}{2 f_{\text{lo}} \sqrt{\varepsilon_r + \frac{1}{2}}} \tag{1}
\]
where \( c \) is the velocity of light in free space. Note that the length and width of the antenna structure, according to (1), are equal to half of the effective wavelength.

Diameter of the half circle representing the ground plane and the radiator is chosen to be equal to \( \frac{w}{2} \). There is a gap equal to \( s \) between the radiating element at the top layer and the ground plane at the bottom layer. Parametric analysis concerning the choice of the best value of \( s \) indicates that it should not be larger than thickness of the substrate \( h \) in order to get the largest operational bandwidth.

Note that the above choice for dimensions of the radiator and the ground plane results in a horizontal tapered dipole of a length equal to a half wavelength, which is elevated from the ground by about a quarter wavelength at \( f_o \).

The above outlined design procedure results in an antenna which covers the whole UWB range from 3.1 GHz to 10.6 GHz. In order to reject a certain band, which extends from a low frequency \( f_1 \) to a high frequency \( f_2 \) within the specified passband, the use of printed parasitic elements with tapered ends is postulated. Such parasitic elements with an outer length \( l_1 \) and an inner length \( l_2 \) are proposed to be placed in the top and bottom layer facing the tapered slot, which is formed by the radiator and the ground plane, as shown in Fig. 1. It is known from the antenna theory that a non-resonating dipole is transparent to incident waves. However, it becomes a strong scatterer when its effective length becomes a half-wavelength at the frequency of the incident wave. This property is applied here to obtain certain frequencies rejection by placing strip dipoles in the radiating aperture of the original UWB antenna.

Following this principle, the parasitic element’s dimensions are selected according to the following expressions:

\[
I_1 = \frac{c}{2f_1 \sqrt{\frac{v}{w} + 1}}
\]

\[
I_2 = \frac{c}{2f_2 \sqrt{\frac{v}{w} + 1}}
\]

If the designed UWB antenna was required to reject a number of discrete narrow bands within the passband, a set of narrow width strips equal to the number of the rejected bands could be used. In this case, length of each of the strips of the parasitic elements would be calculated using (2).

### III. RESULTS

The design of the proposed UWB antenna is undertaken assuming Rogers RO4003C substrate with a dielectric constant equal to 3.38, tangent loss = 0.0027, and thickness \( h = 0.520 \text{ mm} \). The design follows the described guidelines followed by the optimization with the software HFSSv10.

The first design concerns a UWB antenna without a rejection band. Values of the design parameters shown in Fig. 1 calculated using the presented method and optimized using HFSSv10 are \( w = 25 \text{ mm}, l = 25.5 \text{ mm}, w_f = 1.38 \text{ mm}, s = 0.5 \text{ mm} \).

The next design concerns a UWB antenna with a selected rejection band. In this case, parasitic elements are added to the UWB antenna designed in the previous step. In order to verify accuracy of the proposed formula (2) for working out the length of the parasitic element, its length is varied and the rejected frequency is calculated using (2). In this case, the parasitic element width is chosen to be 0.4 mm. The obtained results are compared with those produced by the full-wave electromagnetic simulation HFSSv10. The comparison, which is shown in Fig. 2, indicates that expression (2) gives quite an accurate estimation for length of the required parasitic element.

The next step concerns the design of three UWB antennas with certain rejection characteristics. One of the antennas aims at rejecting a certain narrow band (5 to 6 GHz), the other one is designed to reject a wide band (4 GHz to 7 GHz), whereas the third one is designed to reject three bands (4.5 GHz to 5.5 GHz, 6.5 GHz to 7.5 GHz, and 8.5 GHz to 9.5 GHz). The procedure adopted in the design was to calculate the required dimensions using the proposed method and then the software HFSS was used to optimize the dimensions for the best possible performance. The calculated dimensions of the parasitic elements for the narrowband rejection are: \( l_1 = 20 \text{ mm}, l_2 = 17 \text{ mm} \), whereas the optimized values are: \( l_1 = 18.5 \text{ mm}, l_2 = 16.5 \text{ mm} \). With respect to the wideband rejection design, the calculated dimensions are: \( l_1 = 25 \text{ mm}, l_2 = 14.5 \text{ mm} \), whereas the optimized values are: \( l_1 = 22 \text{ mm}, l_2 = 13 \text{ mm} \). The calculated dimensions for the three-band rejection design are: \( l_1 = 20 \text{ mm}, l_2 = 14.5 \text{ mm}, l_3 = 11.3 \text{ mm} \), whereas the optimized values are: \( l_1 = 18.5 \text{ mm}, l_2 = 14 \text{ mm}, l_3 = 10.6 \text{ mm} \). It is clear that the presented method gives an accurate estimate of the required parasitic lengths where the difference between the calculated and the optimized values is around 10%.

In the next step, the four designed antennas were manufactured and experimentally tested. Characteristics of the developed antennas were obtained using a vector network analyser in an anechoic chamber; whereas the simulations were carried out using HFSSv10. Fig. 3 shows variation of the voltage standing wave ratio (VSWR) with frequency for the developed antennas. The measured and simulated characteristics of the antenna without parasitic elements reveal UWB behaviour with bandwidth from 3 GHz to more than 11 GHz assuming VSWR = 2 (or return loss of 10 dB) as a reference. Concerning the antennas with parasitic elements, Fig. 3 shows that the undesired band is tuned out, whereas the wideband behaviour of the antenna is maintained. VSWR is more than 30 at the center of the rejected bands indicating a complete tuning out of that frequency.

From the UWB applications point of view, the UWB antennas are usually required to have an omnidirectional radiation. Concerning the designed antennas, this requirement is fulfilled over the whole passband of 3.1–10.6 GHz, as shown in the measured radiation pattern in Fig. 4 at three frequencies (3, 6 and 9 GHz) for the two principal planes (xz-plane and yz-plane). The antenna was assumed to be in the yz-plane with the width \( w \) extending along the x-axis. The results, which are shown in Fig. 4, indicate an omnidirectional performance, especially in the xz-plane, knowing that the radiated field was found to be linearly polarized in the y direction. The radiation patterns of the antenna...
with parasitic elements across their passbands are similar to the results shown in Fig. 4 and therefore, they are not presented here.

Fig. 5 shows variation of the measured gain of the developed antennas across the ultra wideband. It is clear from the results in Fig. 5 that the antennas have a low gain, which is expected for omnidirectional antennas. The gain of the antenna without parasitic elements increases with frequency from $-0.2$ dBi at 3 GHz to 3.9 dBi at 9 GHz, and then it decreases a little and becomes 3.6 dBi at 11 GHz. Concerning gain of the developed antennas with parasitic elements, the measured results, presented in Fig. 5, show that the general behavior of the gain is similar to that of the antenna without parasitic elements except for the rejected bands. Across the rejection bands, the gain can be as low as $-8$ dBi, when the parasitic elements are used, compared with $2-3$ dBi for the antenna without parasitic elements. Therefore the antenna gain is suppressed by more than 10 dB. This proves the high capability of the added parasitic elements to reject the undesired bands.

The last test concerns the ability of the designed UWB antennas to transmit and receive pulses. In this case, only the antenna without the parasitic elements is tested. Two identical originally designed UWB antennas without parasitics are used to measure the transmission coefficient between their feeding ports in the frequency domain. In the next step, these results are transformed (via an Inverse Fast Fourier Transform, IFFT) to the time domain using the time-domain capability of HP8510C/HP8530 VNA. Using this in-built function, the frequency range of 3.1 to 10.6 GHz with 100 steps is selected as the basis for carrying out IFFT. The time-domain results for the transmission coefficient obtained when the two co-polarized antennas are separated by a distance of 40 cm are shown in Fig. 6. In this figure, the received pulse is scaled so that its peak value is equal to that of the transmitted pulse. From Fig. 6, it can be observed that the original UWB antenna without the parasitic strips supports an almost distortionless pulse transmission.

**IV. CONCLUSION**

In this communication, parasitic elements in the form of printed strips placed in the radiating aperture of a planar monopole at the top
and bottom layer have been used to reject certain bands within its UWB passband. The radiating element and the ground plane of the adopted antenna's configuration are of a half circle shape. Four samples of antennas have been designed and tested: one without parasitics, the second one with parasitics to reject a certain narrow band, the third with wide parasitics to reject a wide band, whereas the fourth antenna with three narrow parasitics to reject three narrow bands. The results of simulation and measurements presented in this communication have shown that the designed antennas have a bandwidth from 3 GHz to 11 GHz excluding the rejected band. The results have revealed that the undesired bands used by other wireless applications can be excluded from the response of the antenna, as the presence of parasitic elements reduces the antenna gain by 10 dB in these bands.

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A Directive Resonator Antenna Using Degenerate Band Edge Crystals

Salih Yarga, Kubilay Sertel, and John L. Volakis

Abstract—A new small antenna formed by a periodic assembly of two-tone dielectrics (Barium-titanate and Alumina) emulating an anisotropic medium is presented. Directive radiation characteristics are achieved when operated close to the band edge frequencies of a class of anisotropic photonic crystals supporting degenerate band edge (DBE) modes. Unique aspects of the antenna design are its smaller size ($0.79 \lambda_d \times 0.80 \lambda_d \times 0.28 \lambda_d$) and nearly optimum aperture efficiency. The subject 3-D assembly, fed by a slot on a ground plane, achieved a directivity of 10.17 dBi. This communication presents the design of the 3-D assembly using full-wave simulations and provides experimental verification of the overall antenna performance.

Index Terms—Artificial anisotropic dielectrics, degenerate band edge crystals, directive antennas, electromagnetic band gap structures, periodic structures, photonic crystals.

I. INTRODUCTION

Gain enhancement methods for small antennas embedded in material layers have recently been reconsidered in relation to metamaterial applications. Earlier designs [1], [2] consisted of dielectric layers of high permittivity ($\varepsilon \gg 1$) placed over a ground plane (at a distance $\lambda_d/2$ for broadband radiation). By properly tuning the excitation placement and layer thicknesses, substantial gain improvements were achieved. However, these designs required comparatively large structures for practical applications. Specifically, an overall thickness of more than $\lambda_d/2$ is required to realize the essential reflections from the ground plane. Also, these longitudinal resonances imply leaky wave radiation with aperture efficiencies of less than 50%.

Alternatively, periodic superstrates may be used as band-pass filters to realize Fabry–Perot (F-P) resonances within the pass-band of EBGs [3], [4]. These configurations are typically designed to operate at the F-P resonance closest to the first band edge frequency, and thus leveraging the increased (directional) selectivity at the band edge (implying higher Q resonances). Recently introduced degenerate band edge (DBE) crystals [5] further increase the directional selectivity by achieving a maximally flat band edge. Unlike a regular band edge in conventional EBGs ($\partial \omega / \partial K = 0$), the DBE resonance in anisotropic EBGs is characterized by $\partial \omega / \partial K = 0$, $\partial^2 \omega / \partial K^2 = 0$, and $\partial^3 \omega / \partial K^3 = 0$. As a result, the F-P resonances of DBE crystals are much stronger (when compared to ordinary EBGs) [5], [6]. Hence, resonances with quality factors similar to those of ordinary EBGs can be realized using fewer unit cells (i.e., thinner overall dimensions) [7].

Degenerate band edge (DBE) crystals can be realized via uniaxial dielectrics with in-plane anisotropy to yield a maximally flat $k-\omega$ diagram (for plane waves propagating normal to the unit cell). As noted in
DESIGN OF DUAL-BAND MICROSTRIP REFLECTARRAY USING SINGLE LAYER MULTIRESONANCE DOUBLE CROSS ELEMENTS

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Abstract—A multiresonance double cross element is used to design a dual-band reflectarray with dual linear polarization. The proposed element has a single conductive layer structure which makes it easy to manufacture. The results presented in this paper show that the mutual effect between the elements of the two bands is negligible. Hence, it is easy to achieve the phase compensation for each band separately. The simulated and measured results for an element designed to cover the X- and K-bands have confirmed the suitability of the proposed element to build a dual-band reflectarray.

1. INTRODUCTION

The microstrip reflectarray is an antenna that consists of a flat reflecting surface with many microstrip elements and a feed antenna. It uses a suitable phasing scheme to convert a spherical wave produced by its feed into a plane wave [1–6]. The microstrip reflectarray is a high gain antenna which evolved as an efficient and cost-effective replacement of the parabolic reflectors and phased arrays: The parabolic reflector lacks the ability to achieve wide angle beam scanning, whereas the high gain phased array with electronic scanning is very expensive due to its complicated beamforming network and amplifier modules [1].

Some applications have emerged recently; where it is required to design a reflectarray within a limited certain space to cover two widely separated bands, such as the X- and K-bands for NASA space systems [1]. The conventional design of the reflectarray cannot
accomplish the requirements of such dual-band applications. Hence, new methods have been proposed by many authors to design a reflectarray which covers the two bands with a high gain and wide scanning angle capability. A stacked structure which is formed from multiple small square loops at the top layer and a large square loop at the bottom layer were proposed in [7] to achieve a dual band performance with dual linear polarisation; whereas variable size crossed dipoles were presented in [8]. For the case of a closely spaced dual band operation, square loop elements were suggested in [9]. The phase compensation in this case was achieved by using a variable angle rotation technique. For linear polarisation, variable size pairs of dipoles were used for the case of widely or closely separated dual bands [1].

In another important development, the stacked approach was used as a suitable solution to the requirement of dual band operation accompanied by a compact size [10]. Two stacked patches with variable size were used independently for the phase compensation at the two bands. In another multi-layer configuration, perforated patches loaded by slots at the ground plane are used as the radiating elements at C-band and rectangular patches directly loaded by slots are used at K-band [11]. In a recent design [12], a single-layer dual closely separated bands (12 GHz and 14 GHz) orthogonal polarisation reflectarray antenna composed of a combination of split cross and rectangle rings for one band and double split square rings for the other band was proposed. A similar combination was also proposed for a broadband single band operation [13].

In this paper, a single-layer multi-resonance double cross reflectarray element, which was presented in [14], is modified to achieve the dual band operation with a dual linear polarisation. The curved multiresonance cross structure utilized in this paper has a broad bandwidth compared with the single-resonance elements, such as the printed dipoles or patches [15], and it is easy to manufacture compared with the stacked elements. In the presented results, it is shown that the proposed element can operate efficiently at the dual bands 10 GHz and 18 GHz with negligible mutual effect between them.

2. DESIGN

To design a dual band reflectarray, a multi-resonance double cross-element shown in Fig. 1 is considered [8]. The microstrip reflectarray was designed to operate in the X- and K-bands. The reflectarray is assumed to be formed by many of the elements shown in Fig. 1 arranged in a square lattice with periodicity of 15 mm, which is equivalent to half a wavelength at centre of the lower band (X-band), i.e., 10 GHz. They
are assumed to have a double symmetry as required in dual polarised applications. It is to be noted that the chosen value for the cell size prevents the appearance of grating lobes at the higher band, which is 18 GHz in this case, as the inter-element separation is less than one free space wavelength.

The configuration of the chosen element and substrate is shown in Fig. 1. Lengths of the dual cross elements \( (L_1, L_2) \) were changed to show their effect on the phase performance at the two assigned bands, while their widths \( (W_1, W_2) \) were fixed at 0.3 mm. As a general rule, values of the lengths \( L_1 \) (and \( L_2 \)) should vary between quarter and half of the effective wavelength at the lower (and higher) bands in order to achieve the required \( 360^\circ \) phase variation across each of the two bands.

The substrate used to support the cross elements is assumed to consist of a thin laminate of Rogers RT5880 with \( \varepsilon_r = 2.2 \), and thickness \( h = 0.13 \text{ mm} \), in addition to a 6 mm of Foam with a dielectric constant equal to 1.07. The parametric analysis using the software CST Microwave Studio has proven that this combination gives a suitable balance between the required volume occupied by the structure and the phase performance concerning the slope and range.

3. RESULTS AND DISCUSSIONS

Variation of the return loss’s phase was studied as a function of frequency. Only the case of a linearly polarised TEM plane wave, which is normally incident on an infinite periodic array of identical elements, is considered. In this case, the side walls of the equivalent TEM waveguide are formed by a perfect magnetic conductor, while its bottom and top walls are composed of a perfect electric conductor. Using the equivalent unit cell waveguide approach, phase of the reflected wave was calculated for the loaded waveguide. The structure
Figure 2. Variation of phase of the return loss with frequency for different lengths of the two cross elements.

was modelled using the software CST Microwave Studio.

Figure 2 shows variation of phase of the reflection coefficient with frequency for different lengths of the two cross elements. It is clear from this figure that the utilised structure has two resonant frequencies: one at around 10 GHz, while the other is around 18 GHz. Fig. 2 also reveals that the phase range for each of the two resonators exceeds the required $360^\circ$. Effect of varying length of each element on value of the resonant frequency is also shown in Fig. 2. Increasing length of the low-band element $L_1$ from 10 mm to 11.5 mm shifts the first resonant frequency from 10 GHz to 9.5 GHz, while changing length of the high-band element $L_2$ from 5 mm to 6 mm shifts the second resonant frequency from 18 GHz to 17.5 GHz. It is also clear from Fig. 2, that changing length of the high-band element has no effect on the low resonant frequency, and similarly changing length of the low-band element has no effect on the high resonant frequency. This means that it is possible to achieve the required phase compensation for each of the two bands independently by changing length of that band’s element.

To make sure that the low-band element has a negligible effect on the phase performance at the high band, the simulation was carried out for two cases; the first case is when the low-band element has length = 10 mm, while the second case is when there is no low-band element, i.e., $L_1 = 0$. The result, which is depicted in Fig. 3, reveals that the phase performance and value of the high band resonant
frequency is almost constant with or without the presence of the low-band element. Similarly, the simulation was also performed to make sure that the high-band element has negligible effect on the low-band performance. The result shown in Fig. 4 confirms the design

**Figure 3.** Effect of the low-band element on the phase performance of the high-band element.

**Figure 4.** Effect of the high-band element on the phase performance of the low-band element.
expectation that there is no mutual effect between the two elements.

As another step to test the coupling effect on the performance of the two elements that form the double cross cell, the phase performance at the two resonant frequencies 10 GHz and 18 GHz for different lengths of the two multiresonant elements is simulated. The result is shown in Fig. 5 for \( L_1 \) from 7 mm to 12.5 mm with \( L_2 = 5 \) mm, and for \( L_2 \) from 3 mm to 7 mm for \( L_1 = 10 \) mm. It is obvious from Fig. 5 that the two

**Figure 5.** The phase performance of the proposed unit cell at 10 GHz and 18 GHz as a function of the element lengths (\( L_1 \) and \( L_2 \)).

**Figure 6.** The measured and simulated performance of the proposed unit cell.
elements operate almost independently at 10 GHz and 18 GHz.

As a final step in checking performance of the proposed reflectarray, a unit cell with $L_1 = 10 \text{ mm}$, and $L_2 = 5 \text{ mm}$ and a double layer substrate (RT5880 with $h = 0.13 \text{ mm}$, in addition to 6 mm of Foam) was manufactured, and tested using the waveguide approach [1]. Performance of the manufactured cell is shown in Fig. 6. It is clear that the developed cell has two resonant frequencies, which are 11 GHz and 17 GHz according to the measured results, and 10 GHz and 18 GHz according to the simulations. The total phase variation across the two bands is around 800°, which is more than the minimum value (720°) needed for a dual-band operation. Amplitude of the return loss across the band 8 GHz to 20 GHz was also simulated and measured. The measured results shown in Fig. 6 reveal that while the return loss is as low as 0.2 dB across most of the investigated band, it has higher values (more than 0.4 dB) at the resonant frequencies. This result is consistent with the simulated results shown in Fig. 6 and with the previously published findings, which show that the maximum return loss of the reflectarray occurs at its resonant frequencies [16].

4. CONCLUSION

A single-layer multiresonance curved double cross element, which can be used to build a dual-band reflectarray with dual linear polarization, has been presented. The results presented in this paper have shown that the mutual effect between the cross elements of the two bands is negligible, which makes it easy to achieve the phase compensation for each band separately. The simulated and measured results for an element designed to operate at the X- and K-bands have confirmed the suitability of the proposed multiresonance cross element for the design of a dual-band reflectarray.

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Planar Multiband Antenna for Compact Mobile Transceivers

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Abstract—A compact planar antenna for portable multistandard transceivers is presented. The proposed microstrip-fed antenna includes a symmetrical double G-shaped radiator and slotted ground plane. A return loss of better than 10 dB is achieved at the frequency bands PCS (1850–1990 MHz), WLAN+ Bluetooth (2400–2480 MHz), WiMAX (2500–2690 MHz), WiMAX (3400–3600 MHz), HIPERLAN2 (5150–5350/5470–5725 MHz), and IEEE 802.11a (5150–5350–5725–5825 MHz). Moreover, the return loss is more than 6 dB across the DCS band (1.71–1.88 GHz). The proposed antenna is printed on a single-layered FR4 substrate, and it occupies a small volume of $40 \times 30 \times 1.6$ mm$^3$. The simulated and measured performance of the antenna confirms its multiband operation and omnidirectional radiation pattern.

Index Terms—Multiband antenna, portable transceiver, WiMAX, WLAN.

I. INTRODUCTION

T HE RAPID growth in the applications of mobile communication systems means that many functions are to be integrated into a mobile handset. A single handset is now required to deal with multistandard services such as voice, data, video, broadcasting, and digital multimedia. This development has led to a great demand for compact multiband antennas designed specifically to handle multistandard services.

The antennas to be used in mobile handsets should have many features such as ease of fabrication, high-efficiency multiband behavior, low profile, simple structure, low cost, and ease of integration with RF front end. Various antenna configurations are shown to be promising candidates for mobile handsets [1]–[8]. The planar inverted-F antenna (PIFA) presented in [1] has narrow bandwidth that prevents its usage in the modern trend of multiband devices. Several methods were proposed to broaden the bandwidth of the PIFA and make it a multiband planar antenna. In one approach [2], [3], the radiating element is modified by providing several radiating branches and elongating the radiator’s dimension to generate multiple resonant modes.

This approach is utilized to design multiband handset antennas at GSM900, DCS, PCS, and UMTS. The other used technique is through a better utilization of the ground plane [4]–[8]. The problem with some of the adopted techniques is the distribution of slots across the ground, which limits the capability to integrate the antenna with the RF circuitry.

In [9], a multiband coplanar inverted-F antenna was proposed. In the presented structure, a microstrip line that is used to feed the primary radiator is coupled to open-ended slots in the finite-size ground. This arrangement enables the launching of additional resonant frequencies. The proposed configuration is employed to build antennas for the two standards, WLAN and WiMAX.

A printed-loop antenna with wideband characteristics is presented for laptop computer applications [10]. The utilized rectangular loop pattern generates four resonant modes below 4 GHz to support several standards with better than 6 dB return loss.

A planar multiband antenna that comprises a dual-band inverted-F resonator and two parasitic elements is proposed to support six standards [11]. One element of the antenna generates a dipole mode, and another is used to excite a loop mode. The measured results show a return loss of more than 6 dB across the bands of interest.

A folded dual-loop multiband antenna is proposed in [12]. It is fabricated using a pair of symmetric meander strips to form two loops. The design reveals more than 6 dB return loss across the bands GSM/DCS/PCS/UMTS.

In this letter, a simple technique is proposed to design a multiband antenna that covers the standards PCS/WLAN+Bluetooth/WiMAX/HIPERLAN/IEEE 802.11a with more than 10 dB return loss, and DCS with more than 6 dB return loss. The antenna includes a symmetrical double G-shaped radiator that is connected to a microstrip feeder. To achieve additional resonant frequencies and to control the position of those frequencies, an open-circuit slot, short-circuit slot, and pair of narrow slits are embedded in the upper part of the ground plane. Moreover, a short-circuit via is used to connect one arm of the radiator to the slotted ground plane.

II. ANTENNA STRUCTURE

The geometry of the proposed antenna is shown in Fig. 1. It is designed using FR4 substrate (dielectric constant $= 4.4$ and thickness $= 1.6$ mm, loss tangent $= 0.02$). The antenna consists of two inverted G-shaped radiating elements that are symmetrically connected to a microstrip feeder. One of the G-shaped radiators is connected to the ground plane through a short-circuit via. A slotted ground plane is used for the proposed antenna in order to increase the number of resonant
frequencies. The overall dimensions of the antenna is equal to $W \times L = 31 \times 41 \text{ mm}^2$.

In order to show the effect of the short-circuit slot (length $L_{G1}$ and width $W_{G1}$) and the pair of slits ($L_{G2}$ and $L_{G3}$) on the performance of the antenna, the return loss is simulated for three configurations: one without a slot and a pair of slits, another with a slot but without slits, and the last one, which is the final design, includes a slot and a pair of slits. The results using the software CST Microwave Studio are shown in Fig. 2. Adding a short-circuit slot in the ground plane as depicted in Fig. 1 results in an additional resonance at 2.6 and 5.8 GHz by improving the matching between the 2.6- and 5.8-GHz radiators and the input feeder. If a pair of slits is embedded in the ground plane in the manner shown in Fig. 1, the surface current density launched from the microstrip feeder is concentrated along the upper G-shaped radiator as depicted in Fig. 3(b), whereas it is concentrated at the lower grounded G-shaped radiator at 3.5 GHz as revealed in Fig. 3(c). At the upper frequency band (5.5 GHz), the radiation is mainly from the open-ended stub $L_{G4}$ as shown in Fig. 3(d). This stub is chosen to have a length equal to a quarter of the effective wavelength calculated at 5.8 GHz. It is worth mentioning that the pair of slits ($L_{G2}$ and $L_{G3}$) is located underneath the stub $L_{G4}$ to operate as a tuning element for the 5.5-GHz radiator.
to make the radiation at this band omnidirectional. In order to quantify the matching effect of the different stubs used in the proposed antenna, the input impedance of the effective radiators is calculated using the employed software with and without all the stubs. It is found to be \((4i + 1)\hat{z}\), \((7i - j)\hat{z}\), \((i + 1)\hat{z}\), \((4i + 17)\hat{z}\), and \((42 + j17)\hat{z}\) without the stub for the frequencies 2.6, 3.5, and 5.5 GHz, respectively. By including the effect of the stubs, the input impedance becomes \((7i + j\gamma)\hat{z}\), \(47.4i\hat{z}\), and \((i\lambda + j\gamma)\hat{z}\) with the stubs for the frequencies 2.6, 3.5, and 5.5 GHz, respectively.

As an initial design procedure for the proposed antenna, the length of the effective radiator at each resonant frequency as revealed from Fig. 3 is chosen to be a quarter of the effective wavelength at that frequency. For example, the length of the effective radiator at 1.9 GHz, i.e., \(L_{19} = \frac{\lambda_{19}}{4}\) as revealed from comparing the main concentration of the current density in Fig. 3(a) to the definition of dimensions in Fig. 1, is chosen to be equal to a quarter of the effective wavelength (24 mm). Using CST Microwave Studio, the optimum value for the effective radiator at 1.9 GHz was found to be 27.5 mm. The optimum dimensions (mm) of the other design parameters as obtained using the software CST Microwave Studio are the following: \(L_{P} = 18.5\), \(W_{P} = 9\), \(L_{2} = 7\), \(W_{\lambda\gamma} = 4.5\), \(W_{\lambda\gamma} = 3.5\), \(L_{\lambda\gamma} = 3.5\), \(L_{2} = 9\), \(L_{3} = 2.5\), \(L_{4} = 8.7\), \(L_{4} = 2.5\), \(P_{3} = 7\), \(P_{2} = 7\), and \(P_{3} = 9\). The width of the two slits is 0.5 mm.

### III. RESULTS AND DISCUSSION

The proposed antenna was manufactured and tested (Fig. 4). The substrate used for the antenna is FR4 with dielectric constant \(\varepsilon_r = 4.4\), loss tangent \(\tan \delta = 0.02\), and thickness \(= 1.6\) mm. The antenna has an overall size of \(40 \times 30\) mm².

The simulated and measured reflection coefficient of the antenna is shown in Fig. 5. Assuming the 10-dB return loss as a reference, the presented results indicate that the antenna has the following bandwidths: 190 MHz (1.80–1.99 GHz), 400 MHz (2.4–2.8 GHz), 400 MHz (3.3–3.7 GHz), and 1000 MHz (5–6 GHz). Thus, the proposed antenna can support the following standards with better than \(-10\) dB reflection coefficient: PCS, 2.4/5.5-GHz WLAN, Bluetooth, 2.5/3.5/5.5-GHz WiMAX/HIPERLAN2/IEEE 802.11a. If a reflection coefficient of \(-6\) dB is used as a reference, the antenna also covers the DCS band (1.71–1.88 GHz). There is a good agreement between the simulated and measured results of Fig. 5 that confirms the practicability of the proposed antenna.

The effect of changing the size of the ground plane on the resonant frequencies of the antenna was investigated. A parametric analysis was employed to find out the effect of ground length \(L_{G}\) on the values of the resonant frequencies. It was noted that there is no significant change in the performance of the antenna when increasing the length from around 36 to 50 mm. However, if the length is made less than 36 mm, the resonant frequencies, especially at the lower band, start to shift slightly. This means that it is possible to increase the length beyond the assumed value (40 mm) without affecting the resonant frequencies. This is an important feature of the antenna as the performance is not affected by any increase in the length of the ground. Thus, the direct integration of the antenna with other RF devices that have their own ground does not change the antenna’s resonant frequencies.

The maximum gain of the antenna was measured by comparing the received power when using the designed antenna to that received when a reference-gain antenna is used. The measurement was done in an anechoic chamber along the three principal planes. The measured gain of the proposed antenna at different frequencies is depicted in Fig. 6. The maximum gain varies between 0.3 dB at 1.9 GHz and 2.2 dB at 5.5 GHz. Those values for the gain are an indication of the omnidirectional performance of the antenna. In order to quantify the different losses in the antenna, the radiation efficiency of the antenna was calculated using the adopted software. The results shown in Fig. 6 indicate that the antenna has more than 80% efficiency at all of the investigated bands.

The normalized two-dimensional radiation patterns of the proposed antenna measured at 1.9, 3.5, and 5.5 GHz along the three principal planes \((\rho; \phi\lambda, \phi\rho; \lambda\phi, \lambda\rho; \rho\lambda)\) are shown in Fig. 7. The orientation of the antenna with respect to the \(\phi\lambda, \phi\rho\), and \(\lambda\rho\)-axes...
A multiband antenna for portable systems has been presented. The structure of the proposed microstrip-fed antenna comprises mainly a symmetrical double G-shaped radiator. A slotted ground plane is utilized underneath the radiator to generate more resonant frequencies that are required for the modern multistandard mobile systems. The simulated and measured results have shown that the proposed antenna can cover PCS/WLAN/WiMAX/HiPERLAN2/IEEE802.11a with reflection coefficient better than −10 dB, and DCS with reflection coefficient less than −6 dB. The antenna has an omnidirectional radiation pattern in the three principal planes. The compact size of the antenna, which occupies a volume of $41 \times 31 \times 14$ mm$^3$, its multiband frequency coverage, and its omnidirectional properties make it a good candidate for the modern multistandard mobile transceivers.

IV. CONCLUSION

As shown in Fig. 7, the proposed antenna has a dipole-like radiation pattern at the low frequency band with a near-perfect omnidirectional pattern in one plane and an 8-shaped pattern in the other planes. At the high frequency band (5.5 GHz), the antenna has almost omnidirectional properties in all the principal planes.

The compact size, simple structure, omnidirectional properties, and multiband coverage make the proposed antenna an attractive candidate for portable multistandard devices.

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PLANAR MULTIBAND ANTENNA FOR MULTISTANDARD MOBILE HANDSET APPLICATIONS

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ABSTRACT: A planar antenna for multifrequency operation is presented. The proposed microstrip-fed antenna includes a slotted ground plane, a T-shaped radiator, meandered and open circuit strips. A return loss of better than 10 dB is achieved at the frequency bands personal communication system (PCS), (1850–1990 MHz), UMTS (1920–2170 MHz), Bluetooth/WiMAX (2400–2480 MHz), WiBro (2300–2390 MHz), WLAN (2400–2480 MHz), WiMAX (3400–3600 MHz). If a return loss of 6 dB is considered as a standard, the antenna can cover DCS (1710–1880 MHz). The proposed antenna occupies a small volume of $40 \times 30 \times 1.6 \text{ mm}^3$, which makes it attractive to the portable devices. The simulated and measured performance of the antenna confirms its multiband operation and omnidirectional radiation pattern. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 53:2700–2703, 2011; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.26319

Key words: dual-band printed monopole antenna; L-slot; arm-slot; broad forward beam

1. INTRODUCTION

The rapid growth in the applications of mobile communication systems means that a single handset is required nowadays to deal with multistandard services such as voice, data, video, broadcasting, and digital multimedia. This has led to a great demand for designing compact multiband antennas for mobile handsets.

Besides the multiband behavior, the antennas installed in the mobile handsets should have high efficiency, low profile, simple structure, easy fabrication, low cost, and easy integration with the radio frequency front-end. One of the important candidates for mobile handsets is the planar inverted-F antenna (PIFA). Multiple resonant modes are generated in the PIFA by increasing the branches of the radiator, elongating the radiator’s dimension, or modifying the ground plane [1–5]. Some of the proposed methods result in a large antenna size that imposes a practical challenge when embedding the antenna in the limited space available in handheld devices. Most of the proposed antennas [1–3] achieve around 6-dB return loss at the required bands. Moreover, some of the proposed antennas do not achieve the required three-dimensional omnidirectional radiation at all the covered bands.

In this letter, a simple technique is proposed to design a multiband antenna that covers the standards PCS/UMTS/WiBro/WLAN + Bluetooth/WiMAX with more than 10-dB return loss. The microstrip-fed antenna includes a T-shaped radiating element with multiple stubs to generate more resonant frequencies. By properly etching a slot line in the ground plane, the antenna can achieve the required resonant frequencies with omnidirectional radiation in the three principal planes as required in the mobile devices.

Figure 1 Configuration of the proposed antenna. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

2. PROPOSED ANTENNA

The geometry of the proposed antenna is shown in Figure 1. It is designed using FR4 substrate (dielectric constant $= 4.4$ and thickness $= 1.6 \text{ mm}$). The antenna consists of T-shaped radiating elements connected with multiple stubs, one of them strongly coupled to the ground using a short circuit via to generate multiple resonant frequencies. The combination of the T-shaped section and the multiple stubs forms a double F-shaped radiator. To enhance the performance at the resonant frequencies, the return loss becomes better than 10 dB at those resonant frequencies, and to achieve a three-dimensional omnidirectional radiation, slotted ground plane is used. The overall dimensions of the antenna is $W \times L = 30 \times 40 \text{ mm}$.

To show the role of each part of the antenna on the performance, the surface current density of the antenna is shown in Figure 2 at the three resonant frequencies 2.1, 2.3, and 3.5 GHz. The current density is calculated using CST Microwave Studio. At the lowest resonant frequency (2.1 GHz), the surface current density is concentrated along the part of the antenna that resembles an inverted-F structure. Thus, the operation of the antenna at the resonant frequency 2.1 GHz can be compared with that of the PIFA antenna. The top left section of the radiator is folded to reduce the antenna’s dimensions, while maintaining the required resonant trace length. As with any PIFA antenna, section of the antenna introduces capacitance to the input impedance of the antenna, which is compensated by using a short-ended shunt stub. The stub is implemented in the proposed antenna using the lower right section of the radiator, which is connected to the ground plane through a via.

At the next resonant frequency (2.3 GHz), the surface current depicted in Figure 2(b) is concentrated at the upper right section of the radiator that resembles an F-shaped structure. Hence, the performance of the antenna can also be explained in a similar manner to that of the 2.1 GHz. However, the length of the effective radiator in this case is smaller than that required for the 2.1
GHz resonant frequency. For the 3.5-GHz resonant frequency, Figure 2(c) reveals that the excitation is due to the coupling of the microstrip feeder and the short-circuit stub with the ground plane.

The distribution of the surface current density depicted in Figure 2 is used to find the dimensions of the different elements of the antenna. The method adopted in this letter is to make the total length of the path of the surface current density at each resonant frequency equal to a quarter of the effective wavelength calculated at that frequency. The overall structure is then optimized to achieve better than 10-dB return loss across the required frequency bands with an omnidirectional radiation pattern. The optimum dimensions of the different elements of the antenna shown in Figure 1 are: \( W_P = 9 \) mm, \( W_C = 2.5 \) mm, \( W_{DC} = 4.35 \) mm, \( W_{S1} = 3 \) mm, \( L_P = 19 \) mm, \( L_{S1} = 8 \) mm, \( L_{S2} = 5.85 \) mm, and \( L_{S3} = 2.3 \) mm. The width of the two slits is 0.5 mm.

3. RESULTS AND DISCUSSION

The proposed antenna is manufactured using the substrate FR4 with dielectric constant \( \varepsilon_r = 4.4 \) and thickness \( h = 1.6 \) mm. A photo of the manufactured antenna is shown in Figure 3.

The simulated and measured reflection coefficient of the antenna is shown in Figure 4. Assuming the 10-dB return loss as a reference, the presented results indicate that the antenna has
the following bandwidths: 0.9 GHz (1.8–2.7 GHz) and 400 MHz (3.3–3.7 GHz). Thus, the proposed antenna can support the following standards with better than –10 dB reflection coefficient: PCS, UMTS, 2.4-GHz WLAN, Bluetooth, WiBro, 2.5 and 3.5 GHz WiMAX. If a return loss of 6 dB is assumed as a standard, it also covers the DCS band (1.71–1.88 GHz). As shown in Figure 4, there is a good agreement between the simulated and measured results.

The measured gain of the proposed antenna at different frequencies is depicted in Figure 5. The maximum gain varies between 0.2 dBi at 1.8 GHz and 1.4 dBi at 3.5 GHz. Those values for the gain are an indication of the omnidirectional performance of the antenna. To make sure of the level of losses in the designed antenna, the simulated radiation efficiency of the antenna is also shown in Figure 5. It is clear that the antenna has more than 85% radiation efficiency across the investigated frequency bands.

The two-dimensional radiation patterns of the proposed antenna measured at 2.3 and 3.5 GHz along the three principal planes (xz, yz, and xy) are shown in Figure 6. The orientation of the antenna with respect to the x, y, and z axes is clarified in Figure 1. It is worth mentioning that the radiation pattern at 2.1 GHz is similar to that at 2.3 GHz, and thus, it is not shown in Figure 6.

The presented results in Figure 6 indicate omnidirectional characteristics along those planes. The radiation pattern for the copolarized and cross-polarized signals at the three principal planes and different frequencies has no deep nulls in any direction. This is an important factor when choosing antennas for the mobile devices. As those devices operate in a multipath environment, the direction of arrival of the dominant signal varies randomly. If the antenna has a deep null in the direction of the dominant signal, signal dropouts occur. To prevent this undesired situation from happening, the handset antenna should not have deep nulls in any direction. This proposed antenna achieves this target as revealed in Figure 6.

The compact size, simple structure, omnidirectional radiation, and multiband coverage with more than 10-dB return loss should make the presented antenna an attractive candidate for portable multistandard devices.

4. CONCLUSION
A compact multiband antenna for portable systems has been presented. The structure of the proposed microstrip-fed antenna comprises mainly of T-shaped radiator, multiple stubs that form double F-shaped radiator with the T-section, and slotted ground plane. The simulated and measured results have shown that the proposed antenna can cover PCS/UMTS/WiBro/WLAN + Bluetooth/WiMAX with return loss more than 10 dB, and DCS with return loss more than 6 dB. The antenna has an omnidirectional radiation pattern in the three principal planes.
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FUZZY LOGIC-BASED AUTOMATIC GAIN CONTROLLER FOR EDFAs
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ABSTRACT: In this study, the fuzzy logic (FL)-based automatic gain controller (AGC) is designed to obtain a fixed gain level of erbium-doped fiber amplifiers along C band based on experimental results. A FL-AGC with two inputs and one output is considered where the inputs are signal power and signal wavelength and the output is the pump laser current. The output gain variations are examined by varying both input signal levels from 15 to 35 dBm and signal wavelength from 1526 to 1564 nm. The results are well kept within 30-dB gain level with a ripple of ±0.1 by using FL-AGC circuit. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 53:2703–2705, 2011; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.26318

Key words: automatic gain control; fuzzy logic; EDFA; C band

1. INTRODUCTION

Flat gain profile for erbium-doped fiber amplifiers (EDFAs) is an important parameter in wavelength division multiplexing (WDM) and dense WDM applications for long haul optical systems and networks [1–4]. In addition to that the signals that are added/dropped from these systems change the output signal level of each channel producing remarkable optical power transients that propagate throughout the optical network. These power transients can cause serious deterioration of system performance parameters such as output gain, bit-error rate, and signal-to-noise ratio during the transients. For the purpose of overcoming the performance deterioration of optical network, it is critical to keep a fixed optical signal level per channel, and the EDFA gain must be controlled so as to be constant at high speeds. To avoid from these limitations, many automatic gain controller (AGC) schemes for the fixed output gain and decreasing the transients in EDFAs are suggested [5–11].

The AGC schemes are generally classified as either pure electronic, all-optical, or a combination of both optical and electronic feedbacks [5]. In this article, a new AGC scheme using fuzzy logic (FL)-based pure electronic feedback is proposed for controlling the pump current. Electronic control methods are generally based on pump laser power adjustment for stabilizing the output signal level. AGC circuits have typically ~1 dB output gain variation, and they cannot immediately perform the adaptation to different wavelengths and input signal powers. The proposed method is simple, efficient, low cost, and easily carried out, and its output gain ripple is very low, with accurate predictions, which does not require rigorous calculations. There are previously proposed FL models that can be found in literature [12–14]; however, this study deals with the AGC that is applicable to EDFAs operating along C band at any wavelength and signal power.

2. THE STRUCTURE OF THE FL-AGC SYSTEM

In recent years, FL control techniques have been applied to a wide range of systems. The FL approach is very useful in many fields to avoid complicated mathematical equations using expert knowledge and experience by fuzzy rules. FL control is a significant invention for the application of control systems. It has very simple linguistic rules, and it is easy to apply to a system that will be controlled. The proposed FL-AGC deals with the use of FL in the modeling of EDFA to determine the input signal power and wavelength dependence of pump laser current. Figure 1 shows the basic structure of the FL-AGC system.

In Figure 1, input variables, which are signal power and wavelength, are the essential variables to obtain fuzzified data by the fuzzification interface (FI) of FL-AGC. FL unit uses fuzzy sets to operate that are represented by membership functions (MFs) that defines how the input space is mapped to a membership value between 0 and 1, which is shown in Figure 2. In this study, the triangular-shaped MFs is used as it provides both accurate results and fastest calculation time among other MFs used in the analysis. The FL-AGC used in the simulations is Mamdani-type [15] FL-AGC with typical if-then rules structure.

The knowledge base (KB) unit consists of a database (DB) and a rule base (RB) unit. The DB unit provides essential explanations, which are used to explain linguistic control rules (LCRs), fuzzy data manipulation, and the RB that characterizes the control aims and control strategy of the experts by the way of a set of LCRs. The decision-making logic (DML) unit checks the KB to find the output value for the several input values symbolized by the MFs. The communication between DB and DML units is continuous. The RB unit includes a series of fuzzy rules, which defines the relation between input and output variables. In this study, rules are occupied for the signal power and wavelength as shown in Table 1. In this table, the linguistic variables mf1 and mf13 correspond to smallest and biggest MFs, respectively. These rules are used in max–min fuzzy method.

The resulting fuzzy set must be converted to a number that can be sent to the process as a control signal. This operation is called defuzzification, and a defuzzification interface unit is used that performs the reverse process of the fuzzification block. It converts the range of output variables into corresponding universe of discourse and gives nonfuzzy data up from inferred fuzzified data. There are several defuzzification methods. In this

![Figure 1](image-url) The basic structure of the FL-AGC system.
Design of compact planar ultrawideband antenna with dual-notched bands using slotted square patch and pi-shaped conductor-backed plane

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Abstract: A planar antenna with ultrawideband (UWB) performance and dual band-notched characteristics is proposed. The main features of the antenna are the compact dimensions and omnidirectional radiation across the whole band of operation. The radiator of the antenna is a slotted square patch. The ground plane is located at the bottom layer, which also includes a Π-shaped conductor-backed plane used to widen the impedance bandwidth. Dual band-notched characteristics are achieved by an inverted T-shaped strip inside the slotted radiator and a pair of mirror inverted L-shaped slots at the two sides of the radiator. The measured results of the manufactured (12 × 16 mm) antenna on 1.6 mm FR4 substrate show that the antenna operates with voltage standing wave ratio (VSWR) less than two over the frequency band from 2.5 to 10.8 GHz. That wideband is featured by the existence of two notched bands (3.23–4.3 and 5–6 GHz), where the VSWR is more than nine, aimed at suppressing any interference from IEEE802.16 WiMAX (3.3–3.6 GHz), C-band systems (3.7–4.2 GHz) and IEEE802.11a WLAN (5.15–5.825 GHz). The antenna has an omnidirectional radiation across the whole UWB as validated by the measured radiation pattern and gain.

1 Introduction

The ultrawideband (UWB) technology has attracted a huge interest recently because of its unlimited applications in short-range wireless communication, localisation and tracking, medical imaging and monitoring and many more [1–4].

One of the key elements to secure a successful UWB system is an UWB antenna with compact dimensions, proper characteristics and immunity to interferences from nearby systems that use parts of the UWB spectrum.

The main parameters in designing UWB antennas, especially for indoor applications, are easy to manufacture structure, compact size and omnidirectional radiation pattern across the band from 3.1 to 10.6 GHz [1–8]. Since there are several existing systems operating within the UWB frequency spectrum, such as IEEE802.16 WiMAX (3.3–3.6 GHz), C-band system (3.7–4.2 GHz) and IEEE802.11a WLAN (5.15–5.825 GHz), the UWB antenna is required to have the capability to notch those bands, and thus, to protect the UWB system, and those systems at the same time, from any interference between them.

Many designs are available in the literature concerning the UWB antenna with band-notched characteristics. Those designs use different types of slots, slits and parasitic elements in the radiator, the ground plane or even in the feeder to achieve the required band-notching characteristics with limited impact on the required passband [9–23]. Many of the proposed designs in the literature are a result of a trial-and-error approach, whereas there are a few that follow a systematic design approach [17, 19, 23].

In this paper, the target is to present a compact structure with a step-by-step design procedure. The final performance of the antenna is aimed at achieving the required UWB and to have dual notched bands that can be adjusted using an empirical formula. The first notched frequency band is achieved by using a pair of mirror inverted L-shaped slots embedded in the radiator, whereas the second notched band is realised by an inverted T-shaped strip inside the radiator. The impedance bandwidth is enhanced by using a Π-shaped conductor-backed plane. The antenna has a compact size of 12 × 16 mm on 1.6 mm FR4 substrate. The presented design is validated by simulations and measurements.

2 Antenna design

Assume that a substrate with a dielectric constant of \( \epsilon_r \) is chosen to support an antenna that operates across, at least, the UWB frequency range from 3.1 to 10.6 GHz. The centre frequency \( f_c \) of the band in this case is 6.85 GHz. The design suggested in this paper starts by assuming that a square path of dimensions \( A / 2 \times A / 2 \) (where \( A \) is the guide wavelength = \( c / \sqrt{\epsilon_r} f_c \), and \( c \) is the speed of light in free space) is placed at the top layer of a substrate as
The antenna covers the band from 3.4 to 10.5 GHz with VSWR less than two.

To modify the performance of the antenna by creating two notched sub-bands at the WiMAX (3.3–3.6 GHz), C-band (3.7–4.2 GHz) and WLAN (5.15–5.825), the square patch is slotted in the manner shown in Fig. 1b. A pair of mirror-inverted L-shaped slots at the two sides of the radiator is created so that the first notched band is centred at 3.75 GHz, whereas the inverted T-shaped strip inside the radiator is responsible for making the second notched band centred at 5.48 GHz. The slot’s length $L_{p1}$ defines the first notched band, whereas the strip’s length $L_{p2}$ defines the second notched band. The relation between the centre of the notched bands ($f_{p1}$ and $f_{p2}$) and those two design parameters is

$$L_{p1} = \frac{c}{4f_{p1} \sqrt{\varepsilon_r}}$$

$$L_{p2} = \frac{c}{4f_{p2} \sqrt{\varepsilon_r}}$$

$L_{p1}$ is the effective dielectric constant, which can be calculated for the microstrip structures using the formula in [24]. For the notched bands that are centred at the frequencies 3.75 and 5.48 GHz, the values of the designed parameters $L_{p1}$ and $L_{p2}$ can be calculated from (1) and (2) as 11.2 and 7.6 mm, respectively. Using the calculated values for $L_{p1}$ and $L_{p2}$, the overall dimensions of the antenna is optimised by using HFSS. The optimised values are shown in Fig. 1b. The performance of the antenna is depicted in Fig. 2 concerning the VSWR. It is clear that the antenna covers the band from 2.5 to 9.8 GHz with VSWR that is less than two. The two bands centred at 3.75 and 5.48 GHz are notched with VSWR that is larger than 20 for the first band and nine for the second band. However, it is clear also from Fig. 2 that the antenna has a poor performance at the frequency band between 9.8 and 10.6 GHz, which should be part of the UWB spectrum. To improve the performance at that band, a II-shaped conductor-backed structure is included at the bottom layer symmetrically oriented with respect to the longitudinal direction of the antenna as depicted in Fig. 1c. The dimensions of the added structure are optimised for the best possible performance at the frequencies 9.8 GHz and above without negatively

---

**Fig. 1** Antenna design  
(a) Configuration of an initial design (square patch antenna)  
(b) Top layer of the final design (units, mm)  
(c) Bottom layer  
Optimised dimensions are shown in (b) and (c)

---

**Fig. 2** Simulated VSWR for the initial square patch structure, the slotted radiator structure and the slotted radiator with backed conductor

**Fig. 3** Simulated VSWR of the antenna with different values for $L_{p1}$
affecting the performance at the lower band. The optimised dimensions of the \(P\)-shaped conductor are shown in Fig. 1c. The final performance is shown in Fig. 2. It is clear that the performance has been improved significantly at the band between 9 and 11.7 GHz without affecting the performance at the lower band. The final design covers the band from 2.5 to 11.7 GHz using \(VSWR = 2\) as the reference.

In order to verify the validity of the design (1) and (2), the performance of the antenna for different values of \(L_{p1}\) and \(L_{p2}\) is calculated using the simulation tool and the results are shown in Figs. 3 and 4. It is clear that the design parameter \(L_{p1}\) defines the position of the first notched band, whereas the second design parameter \(L_{p2}\) defines the second notched band. It is possible to show using the design equation of microstrip structures [24] that for the utilised structure, the effective dielectric constant is given approximately as \(\varepsilon_{re} = 3.2\). If this value is substituted in (1) and (2) along with the values of \(L_{p1}\) and \(L_{p2}\) that are used to generate the simulation results of Figs. 3 and 4, the location of the centre of the notched bands \(f_{p1}\) and \(f_{p1}\) calculated from (1) and (2) are almost the same simulated values shown in Figs. 3 and 4.

In another step to obtain some insight into the physical meaning of the generated notched bands by the utilised L-shaped slots and T-shaped strip, the surface current distribution at the centre of the two notched bands and a frequency within the passband are calculated using HFSS and is shown in Fig. 5.

It is clear from Fig. 5a that the current flows in opposite directions at the two edges of the inverted L-shaped slots at 3.7 GHz. Thus, the total effective radiation is very low, and thus a notched band is achieved. In Fig. 5b, the surface current at 5.4 GHz at the internal inverted T-shaped strip is in reverse direction to the current in the outer edges of the radiator. Thus, the overall radiation at this band is very limited and a second notched band is achieved. Outside the notched bands, the whole structure behaves as a radiator as indicated by the simulated current distribution of the
antenna at 9.5 GHz (Fig. 5c). The current at the different parts of the antenna are in the same direction, and thus an effective radiation occurs.

3 Measured results and discussion

The developed antenna that has the dimensions $W_{\text{sub}} = 12\text{ mm}$ and $L_{\text{sub}} = 16\text{ mm}$ was tested using an Agilent 8722ES Vector Network Analyzer (VNA). The main problem facing the accurate measurements of the characteristics of the designed antenna, just like any other small antenna, is the effect of the coaxial cable connecting the antenna with the VNA. If not decoupled properly, the measurement cable becomes part of the radiating structure, and thus, changes the input impedance of the antenna. Moreover, the cable distorts the far-field radiation pattern of the antenna by reradiating part of the signal that leaks into the cable. For those reasons, several techniques were proposed in the literature to decouple the cable from the antenna, such as using different types of baluns, ferrite beads or optic links [25–27]. In this paper, high impedance ferrite beads are employed along the measurement cable close to its connection with the antenna to reflect and/or absorb the induced power on the cable. This action minimises the effect of the cable on the antenna significantly.

The results concerning the VSWR values are shown in Fig. 6. They indicate an impedance bandwidth from 2.5 to 10.8 GHz excluding the two notched bands (3.2–4.2 and 5–5.9 GHz). There is generally a good agreement between the simulated and measured results.

The other important parameter needed to verify the performance of the UWB antenna is the variation of the group delay across the band of operation. For a distortion less performance, the deviation in the group delay at the passband should be as small as possible. The group delay was measured for the developed antenna by using two antennas as a transmitter and receiver. The distance between the antennas was 10 cm and they face each other in the broadside direction. The mastered group delay is shown in Fig. 6. It is clear that the peak-to-peak variation in the group delay is less than 0.5 ns in the passband, whereas it is up to 3 ns in the notched bands.

To confirm the omnidirectional behaviour of the antenna, the measured radiation pattern at the frequencies 3, 4.5, 8 and 10 GHz, in the $x$–$z$ and $y$–$z$ planes are shown in Fig. 7. From an overall view of these radiation patterns, the proposed antenna behaves quite similar to the typical printed monopoles in the lower and middle frequency bands. The H-plane patterns are almost omnidirectional, but they are more directional in the higher band.

Fig. 8 shows the measured maximum antenna gain from 3 to 11 GHz for the developed antenna. The simulated gain of a square patch structure is also shown to confirm the effect of the utilised approach in the rejection of two sub-bands. The
A compact microstrip-fed printed monopole antenna with UWB performance and dual band-notched characteristics has been presented. The radiator of the antenna is in the form of a slotted square patch, whereas the ground is a reduced size rectangular plane. The slots dimensions are selected according to an empirical formula to create two notched bands at the desired frequencies. The first notched band aimed at preventing any interference with existing WiMAX and C-band systems is achieved by using a pair of mirror inverted L-shaped slots in the radiation patch, which exempt from interfaces. The second notched band aimed at interfering with the 5 GHz WLAN systems is achieved by an inverted T-shaped strip extended inside the slotted radiator. A II-shaped conductor-backed plane with proper dimensions is used to extend the bandwidth of the antenna. The measured results of a 12×16 mm antenna on an FR4 substrate that has 1.6 mm thickness show a wide impedance bandwidth (131%) from 2.5 to 10.8 GHz, two notched bands centred at 3.7 and 5.4 GHz and an omnidirectional radiation.

4 Conclusion

The realised dual band-notched antenna has good gain flatness except at the two notched bands. As shown in Fig. 8, the gain decreases drastically at the notched bands.

All of the aforementioned characteristics certify that the antenna is a promising candidate for UWB systems that need to avoid any interference with nearby wireless systems using the bands WiMAX (3.3–3.6), C-band (3.7–4.2) and WLAN (5.15–5.825).

5 References

Design of Compact Directional Couplers for UWB Applications

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Abstract—This paper presents a simple design method for a class of compact couplers, which offer coupling in the range of 3–10 dB over an ultra-wide frequency band from 3.1 to 10.6 GHz. The proposed couplers are formed by two elliptically shaped microstrip lines, which are broadband coupled through an elliptically shaped slot. Their design is demonstrated for a 3-, 6-, and 10-dB coupling assuming a 0.508-mm-thick Rogers RO4003C substrate. Results of simulation and measurements show that the designed devices exhibit a coupling of 3±1 dB, 6±1.4 dB and 10±1.5 dB across the 3.1–10.6-GHz band. This ultra-wideband coupling is accompanied by isolation and return loss in the order of 20 dB or better. The manufactured devices including microstrip ports occupy an area of 25 mm × 15 mm.

Index Terms—Compact ultra-wideband (UWB) couplers, coupled circuits, directional couplers, planar coupler design.

I. INTRODUCTION

BROADBAND microwave directional couplers are a very important category of passive microwave circuits. They are used to combine or divide signals with appropriate phase of ±90°, and are commonly used in microwave subsystems such as balanced mixers, modulators, and antenna beam-forming networks [1]. In addition, they are essential for developing the cost-effective measurement equipment [2]–[4]. Our particular interest in these devices is with respect to developing an ultra-wideband (UWB) microwave imaging system for breast cancer detection [5], [6]. In these and many other applications, the required couplers are often required to be accomplished in planar (stripline or microstrip) technology. In order to achieve their broadband operation, the approach of coupled transmission lines can be employed. The inherent feature of this approach is that matching and directivity is perfect, and independent of frequency, at least under ideal conditions. However, the challenge is to obtain a tight coupling in the range of 3–6 dB.

Using coupled microstrip lines, the tight coupling can be accomplished using the Lange [7] or tandem coupler configurations [8]–[10]. However, they require wire crossovers, which is inconvenient from the manufacturing point-of-view. In addition, the Lange coupler features narrow strips, which create additional manufacturing problems due to the requirement for strict etching tolerances. In turn, the broadband tandem coupler may require wiggles or serpentines to equalize even- and odd-mode phase velocities [9], [11] when realized in microstrip technology.

In order to avoid these problems, the slot-coupling approach involving a double-sided substrate, which was first proposed by Tanaka et al. [12], can be applied to realizing a tight coupling. The structure is formed by two microstrip lines separated by a rectangular slot in the common ground plane. Its design formulas were given in [13].

When one aims only at the design of a 3-dB coupler, an alternative is the microstrip-slotline approach, which was described by de Ronde [11]. In contrast to Tanaka et al., the de Ronde’s approach preserves the one-layer microstrip format of the coupler at an expense of etching both sides of a ceramic substrate. One side of this coupler is formed by two parallel connected microstrip lines, while the other one includes a straight slotline with two circular terminating slots. In addition, de Ronde suggested the use of a capacitive disc below the slotline to enhance broadband performance. A very important feature of this coupler is a multioctave operation and a very compact size.

By introducing modifications to the original de Ronde’s design, Garcia [14] demonstrated an alternative configuration of a compact planar 3-dB coupler operating, similarly as de Ronde’s device, over the 4 : 1 bandwidth. In his design, Garcia avoided the circular terminating slots and the capacitive disc. Instead, he enlarged the size of a slot below the microstrip layer. This could be the key to achieving UWB performance.

By neglecting the capacitive disc beneath the slot, which appeared in the original de Ronde’s configuration, Schiek [15], and then Hoffmann and Siegl [16], produced the design rules for the microstrip-slot 3-dB coupler. However, for the simplified configurations, their designs were not as broadband as offered by de Ronde and Garcia.

In this paper, we describe a class of compact planar couplers, which are capable of providing coupling between 3–10 dB over an ultra-wide frequency band. In order to find initial dimensions of these devices, simple design equations similar to the ones described in [13] are applied. Final dimensions are obtained with the use of full-wave electromagnetic analysis software package such as Ansoft’s High Frequency Structure Simulator (HFSS). The validity of the presented designs is confirmed experimentally.
For the equivalent rectangular shaped microstrips and slot, the analysis and design procedure is similar to the one described in [13]. Assuming that the coupler is required to have $\epsilon'_{\text{DB}}$ coupling, the even ($Z_{\text{ee}}$) and odd ($Z_{\text{oo}}$) mode characteristic impedances are calculated using (1) and (2) as follows:

$$Z_{\text{ee}} = Z_0 \left( \frac{1 + \left( \frac{\epsilon'_{\text{DB}}}{2} \right)^2}{1 - \left( \frac{\epsilon'_{\text{DB}}}{2} \right)^2} \right)^{0.5}$$  \hspace{1cm} (1)

$$Z_{\text{oo}} = Z_0 \left( \frac{1 - \left( \frac{\epsilon'_{\text{DB}}}{2} \right)^2}{1 + \left( \frac{\epsilon'_{\text{DB}}}{2} \right)^2} \right)^{0.5}$$  \hspace{1cm} (2)

where $Z_0$ is the characteristic impedance of the microstrip ports of the coupler.

Assuming that $Z_0 = 50\,\Omega$ and the coupling factor $\epsilon'_{\text{DB}}$ is 3, 6, or 10 dB, the values of $Z_{\text{ee}}$ and $Z_{\text{oo}}$ can be calculated from (1) and (2) and are given as follows: 120.5 and 20.7 $\Omega$ for $\epsilon'_{\text{DB}} = 3\,\text{dB}$, 86.7 and 28.8 $\Omega$ for $\epsilon'_{\text{DB}} = 6\,\text{dB}$, and 69.4 and 36.0 $\Omega$ for $\epsilon'_{\text{DB}} = 10\,\text{dB}$.

Before commencing the design, we consider the operation of this coupler for the odd and even modes. When the odd mode is excited, the slot can be replaced by a perfect electric conductor. The resulting upper part of the equivalent coupler shown in Fig. 1(b) becomes a microstrip line whose characteristic impedance is $Z_{\text{oo}}$. The width $w_p$ realizing $Z_{\text{oo}}$ can be determined using standard design equations for a microstrip transmission line [17]. Alternatively, the static formulas described in this paper can be used. From Fig. 1(c), one can see that, in the odd mode, the electric field concentrates mostly in the parallel-plate region formed by the patch and ground plane.

A fringe effect, also observed in Fig. 1(c), is less pronounced for small $Z_{\text{oo}}$ as $w_p$ becomes large in comparison with the substrate thickness $h$.

A different wave propagation condition occurs under the even-mode wave excitation. For this mode, the magnetic conductor replaces the slot in the ground plane. Its presence pushes an electric field (launched from the microstrip port) outside the parallel-plate region. This is because the magnetic conductor forming the lower plate does not allow the electric field to be perpendicular to its surface. As a result, the even-mode wave travels in two antipodal slot regions outside the parallel-plate region, as shown in Fig. 1(c). In order to enable a smooth launch of the even-mode wave from the microstrip port to the two antipodal slotlines, the transition formed by the elliptically shaped patches and the ground slot, as shown in Fig. 1(a), is required.

The dimensions $w_p$ and $l_p$ of the equivalent rectangular shaped coupler [see Fig. 1(b)], providing the required even- and odd-mode characteristic impedances, are determined using a static approach similar to the one presented in [13]. By using this approach, $Z_{\text{ee}}$ and $Z_{\text{oo}}$ are given by (3) and (4) as follows:

$$Z_{\text{ee}} = \frac{\epsilon'_{\text{DB}}}{\sqrt{\mu_r}} \frac{K(l_1)}{K'(l_1)}$$  \hspace{1cm} (3)

$$Z_{\text{oo}} = \frac{\epsilon'_{\text{DB}}}{\sqrt{\epsilon_r}} \frac{K(l_2)}{K'(l_2)}$$  \hspace{1cm} (4)

Fig. 1. (a) Layout of the proposed wideband coupler including microstrip ports. (b) Equivalent configuration used to work out initial dimensions. (c) Electric field lines for odd- and even-mode excitation.
where $K'(l)$ is the first kind elliptical integral and $K''(l) = K(\sqrt{1 - l^2})$. Following [13], the parameters $l_1$ and $l_2$ are calculated using (5) and (6) as follows:

\[
\begin{align*}
    l_1 &= \frac{\sinh^2(\pi v_s / 4l)}{\sqrt{\sinh^2(\pi v_s / 4l) + \cosh^2(\pi v_p / 4l)}} \\
    l_2 &= 1 - l_1
\end{align*}
\]

where $h$ is the thickness of the substrate, $v_p$ is the width of the top and bottom microstrip patches, and $v_s$ is the width of the slot of Fig. 1(b).

Using the analysis in [18], the ratio of elliptical functions appearing in (3) and (4) can be approximated by the following:

\[
\frac{K'(l)}{K''(l)} = \begin{cases} 
    \frac{2}{\pi} \ln \left( \frac{1 + \sqrt{1 - l^2}}{1 - \sqrt{1 - l^2}} \right), & \text{for } 1 < l < 17 \\
    \frac{2\pi}{\pi} \ln \left( \frac{1 + \sqrt{1 - l^2}}{1 - \sqrt{1 - l^2}} \right), & \text{for } 1 < l < 17,
\end{cases}
\]

The synthesis task of determining $v_p$ and $v_s$ for given values of $\varepsilon_{ee}$ and $\varepsilon_{eo}$ is accomplished by solving (3)–(7) using the Gauss–Newton iteration method.

The last step of the design procedure concerns the determination of the coupler’s length. Here, $l_p = l_s = \lambda_e/4$, where $\lambda_e$ is the effective wavelength for the microstrip line and can be calculated using standard formulas such as those presented in [17].

Formulas (3)–(7) enable calculations of the equivalent parameters of the rectangular shaped coupler of Fig. 1(b). The next step is to work out the dimensions of the elliptically shaped counter part. Due to compact size, where the dimension is equal or less than a quarter of the effective wavelength, one can expect a similar performance when the rectangular and elliptically shaped couplers occupy an approximately equal area. Using this equivalence principle and assuming that the mean algebraic length of the elliptically shaped coupler is equal to its rectangular counterpart such that $l_3 = (\sqrt{l_p^2 + v_p^2} + l_p)/2$, then the width of the microstrip $l_1$ and the width of the slot $l_2$ for the elliptically shaped coupler can be obtained using (8) and (9) as follows:

\[
\begin{align*}
    l_1 &\approx \frac{2\pi}{\pi} \ln \left( \frac{1 + \sqrt{1 - l_3^2}}{1 - \sqrt{1 - l_3^2}} \right) \\
    l_2 &\approx \frac{2\pi}{\pi} \ln \left( \frac{1 + \sqrt{1 - l_3^2}}{1 - \sqrt{1 - l_3^2}} \right)
\end{align*}
\]

The final dimensions $l_1$, $l_2$, and $l_3$ are adjusted by iteratively running the finite-element method design and analysis package Ansoft HFSS v9.2.

In order to test the coupler experimentally, its ports need to be connected to SMA coaxial connectors. To minimize possible reflections, curved microstrip lines, as shown in Fig. 1(a), can be used. Our simulations have revealed that for high-quality impedance match, the radius of these curved lines should not be less than twice the width of the microstrip line.

### III. Results

The validity of the presented design method is tested in examples of 3-, 6-, and 10 dB directional couplers aimed for operation in the 3–10-GHz frequency band. For this band, the center frequency of operation is 6.5 GHz. A Rogers RO4003C substrate featuring a dielectric constant of 3.38 and a loss tangent of 0.0027, 0.508-mm thickness, plus 17-μm-thick conductive coating is selected for the couplers development.

Using the proposed method, the dimensions $l_1$, $l_2$, and $l_3$ are determined and are shown in Table I. One can find that the obtained values are not too far off from the ones calculated using (3)–(9). First, $v_p = 3.04$ mm using [13] or $v_p = 4.4$ mm using (4), $l_p = l_s = 6.84$ mm, $v_s = 4.4$ mm, and $l_1 = 1.14$ mm. Therefore, $l_2 = 7.2$ mm (for $v_p = 3.04$ mm), $l_3 = 4.7$ mm, and $l_3 = 7.4$ mm.

The return loss, coupling, and isolation of the designed couplers are first verified using HFSS. Fig. 2 shows the simulated amplitudes of the scattering parameters for the designed 3-dB coupler. These are followed by results of the phase difference between the two output ports, as shown in Fig. 3.

It is clear that the designed coupler features UWB characteristics. The coupling is 3 ± 0.8 dB for the 3.1–10.6 GHz band. The isolation and return loss are better than 28 and 22 dB, respectively, for the band. In Fig. 3, it is observed that the phase difference between ports 2 and 3 is 90° ± 1° over the band. This

### Table I

VALUES OF DESIGN PARAMETERS IN MILLIMETERS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value for 3-dB</th>
<th>Value for 6-dB</th>
<th>Value for 10-dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>$l_1$</td>
<td>120.5</td>
<td>56.7</td>
<td>60.4</td>
</tr>
<tr>
<td>$l_2$</td>
<td>29.7</td>
<td>29.8</td>
<td>29.6</td>
</tr>
<tr>
<td>$l_3$</td>
<td>4.8</td>
<td>4.8</td>
<td>4.8</td>
</tr>
<tr>
<td>$l_1$</td>
<td>7.2</td>
<td>7.2</td>
<td>7.2</td>
</tr>
<tr>
<td>$l_2$</td>
<td>7.2</td>
<td>7.2</td>
<td>7.2</td>
</tr>
</tbody>
</table>

Fig. 2. Simulated performance of the designed 3-dB directional coupler.
result together with the magnitude results shown in Fig. 2 indicates that the coupler operates as a backward wave quadrature coupler [17].

Figs. 4 and 5 show the simulated amplitudes of the scattering parameters for the designed 6- and 10-dB couplers.

It can be seen that, for the 6- and 10-dB couplers, the best result for the coupling is obtained for frequencies around the center frequency. The gradual deviation from the specified value of coupling then occurs. In general, the three couplers feature quite a good UWB performance despite only being formed by a one-quarter-wave section of (nonuniformed) coupled lines.

The directional couplers are then manufactured and tested using a vector network analyzer. The photograph of the one of the manufactured 3-dB couplers is shown in Fig. 6.

The overall dimensions of the coupler including bent microstrip lines are 25 mm × 15 mm, indicating that the device is of a very compact size. The manufactured 6- and 10-dB couplers have the same size.

The measured results are presented in Figs. 7–9. As observed in Figs. 7–9, all of the manufactured couplers show UWB behavior with coupling 3±0.8, 6±1.4, and 10±1.5 dB for the 3-, 6-, and 10-dB couplers, respectively, across the 3.1–10.6-GHz band. The isolation is better than 23, 20, and 19 dB, while the return loss is better than 21, 18, and 19 dB for the 3-, 6-, and 10-dB couplers, respectively.

As observed from the presented data in Figs. 7–9, the operation of the 3-dB coupler seems to be best and is superior over the one of Garcia [14], which showed the 3 dB±1-dB bandwidth from 4.5 to 8 GHz and the isolation of around 20 dB.

The manufactured 6- and 10-dB couplers exhibit some insertion losses, which are not observed in the simulated results. These can be due to conduction and dielectric losses, the difficulty of manual aligning the two microstrip layers forming this type of coupler, and coaxial connectors. The 6- and 10-dB couplers have a smaller width than the 3-dB coupler and as such they are more sensitive to aligning errors. However, in general, the agreement between the simulated and measured results can be considered as very good.
In some applications, one may wish to house the designed couplers in enclosures. In this case, it is important to assess the effect of shielding. Here, this problem was investigated only via computer simulations. Only brief comments concerning the results of these simulations are reported. The produced simulation results revealed that a metal cover with a height of 0.5 cm below and above the three investigated couplers did not adversely affect their performance, as the electrical characteristics were very similar to those shown in Figs. 2–5. Only small adverse effects of the enclosure were observed when the shielding height above and below the coupler structure was reduced to 0.25 cm.

IV. CONCLUSION

A simple method has been proposed for the design of compact directional couplers for UWB applications. The proposed devices are formed by a multilayer microstrip structure with broadside slot coupling. The coupling is controlled by elliptical shapes of microstrip conductors and a coupling slot. The design method has been demonstrated for the case of 3-, 6-, and 10-dB coupling. The couplers have been manufactured and experimentally tested. They have shown UWB behavior across the band from 3.1 to 10.6 GHz. Due to compact size and good electrical performance, they should be of considerable interest to the designers of UWB components. Our particular aim is to use them in a UWB microwave imaging instrumentation [4]–[6].

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and industrial applications of microwaves.

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DESIGN OF ULTRA WIDEBAND 3DB QUADRATURE MICROSTRIP/SLOT COUPLER

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ABSTRACT: This article presents the configuration and design method of a compact ultra wideband 3dB quadrature coupler employing a microstrip/slot technology. The proposed device uses two substrates with a common ground plane. It is formed by two identical multisection elliptical conducting strips, which are broadside coupled through a multisection elliptical or a slot in the ground plane. Results of simulation and measurements show that the coupler exhibits a coupling equal to 3 ± 1 dB for the band 2.3–12.3 GHz. The isolation and return loss are better than 23 dB across that band. © 2007 Wiley Periodicals, Inc. Microwave Opt Technol Lett 49: 2101–2103, 2007; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.22674

Key words: directional coupler; quadrature coupler; UWB coupler

1. INTRODUCTION

A 3 dB microwave directional coupler is a very important passive microwave device, which is used for signal dividing/combining and phasing in many microwave sub-systems [1].

For ease of integrating with other passive or active components it is often required to be designed in planar technology. In order to achieve its broadband operation, an approach of coupled transmission lines is often employed.

Examples of broadband 3 dB planar couplers that use coupled lines include the Lange [2] and tandem couplers [3]. Their shortcoming is that they require wire crossovers, narrow or wiggled strips, which is inconvenient from the manufacturing point of view. Alternative approaches to realizing broadband 3dB planar couplers include those introduced by De Ronde [4], Garcia [5] and Tanaka et al. [6]. The important feature of these alternatives is that they use microstrip/slot technology to achieve a very compact design of a 3dB coupler offering a 4:1 bandwidth by using just a quarter-wavelength section of coupled lines.

By extending the idea of Tanaka et al. [6], we presented in [7] designs of compact planar couplers, which are capable of providing a coupling between 3 and 10 dB across the band 3.1 to 10.6 GHz. To achieve UWB performance we proposed to use elliptically shaped broadside coupled strips and a slot created in a common ground plane of two dielectric substrates. Via full EM computer simulations we found that an elliptical shape provides the best performance of this type of coupler when its electrical length is set at quarter wavelength at the center frequency of the design band. This is one of major differences to the design of Tanaka et al. which used rectangular section of the strips and the slot.

The design presented in [7] uses bent sections of microstrops, which are attached to the elliptically shaped strips to make connections to external ports. The bent microstrip lines offer a possibility of inclusion of new elliptically shaped microstrip/slot sections to extend the operational bandwidth of this coupler. This possibility is explored in the design of a cascaded coupler operating from 2.3 to 12.3GHz, which is described in this article. The validity of the proposed design is confirmed experimentally.

2. DESIGN

The configuration of the proposed directional coupler is shown in Figure 1. The difference between this configuration and the one presented in [7] concerns inclusion of two extra microstrip/slot sections on the left and right hand sides of the middle microstrip/slot section. In contrast to the original design, the two new microstrip/slot sections replace part of the bent sections of microstrip lines that make connections to external ports.

As shown in Figure 1 the proposed coupler consists of three conductor layers interleaved by two dielectrics. The top conductor layer contains the Ports 1 and 2. The bottom conductor layer is identical to the top layer but the ports there are designated as 3 and 4. The two layers are coupled via a slot, which is made in the conductor supporting the top and bottom dielectrics. As seen in Figure 1(a) and (c) the top and bottom conductor layers are formed from the connection of a central ellipse with two side ellipses that are smaller in size.

The analysis and design procedure can be initiated by extending the method presented in [7] to the case of three sections symmetrical coupler [8]. By assuming that the length of each

Figure 1 Layout of the proposed UWB 3dB-coupler. (a) top layer, (b) mid layer, (c) Bottom layer and (d) the whole structure. Diameters of the ellipses in the j-th section are Dj, where j = 1 for primary diameter of the microstrip patches in (a) and (c), j = 2 for primary diameter of slots in (b) and j = 3 for the secondary diameters. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
section is around quarter of the effective wavelength at the center frequency it can be shown that:

\[ C = C_2 - 2C_1 \]  

(1)

where \( C_2 \) and \( C_1 \) are coupling of the central and outer sections, respectively. The even \((Z_{oei})\) and odd \((Z_{ooi})\) mode characteristic impedances for each section are calculated using Eqs. (2), (3):

\[ Z_{oei} = Z_o \frac{1 + C_1}{\sqrt{1 - C_1}} \]  

(2)

\[ Z_{ooi} = Z_o \frac{1 - C_1}{\sqrt{1 + C_1}} \]  

(3)

where \( Z_o \) is the characteristic impedance of the microstrip ports of the coupler.

Assuming that \( Z_o = 50 \ \Omega \) and the coupling factor \( C = 0.708 \) or \( 3 \ dB \), and \( C_2 = 0.892 \) or \( 1 \ dB \) then from Eqs. (1), (2), and (3):

\[ C_1 = 0.092 \ or \ 20.7 \ dB, \ Z_{oe1} = 54.8 \ \Omega, \ Z_{oo1} = 45.6 \ \Omega, \ Z_{oe2} = 209 \ \Omega, \ Z_{oo2} = 12 \ \Omega. \]

The dimensions of the elliptical microstrips and slots offering the required even and odd mode characteristic impedances can be determined using a static approach [6]. Using this approach, \( Z_{oei} \) and \( Z_{ooi} \) are given by:

\[ Z_{oei} = \frac{60\pi}{\sqrt{k_1} \ K(k_1)} \]  

(4)

\[ Z_{ooi} = \frac{60\pi}{\sqrt{k_2} \ K'(k_2)} \]  

(5)

where \( K(k) \) is the first kind elliptical integral and \( K'(k) = K(\sqrt{1 - k^2}) \). The parameters \( k_1 \) and \( k_2 \) have a direct relation with dimensions of the \( i \)-th section of the coupler as shown below.

Rearranging the equations in [7] it is possible to find the design parameters:

\[ k_{1a} = \frac{\sinh^{-1} (0.617D_iD_jlh)}{\sinh^{-1} (0.617D_iD_jlh) + \cosh^{-1} (0.617D_iD_jlh)} \]  

(6)

\[ k_{2a} = \tanh (0.617D_iD_jlh) \]  

(7)

where \( h \) is thickness of the substrate, \( D_i \) and \( D_j \) are diameter of the \( i \)-th section of the elliptical microstrip line and slot, respectively, \( D_{ij} \) is length (or secondary diameter) of the \( i \)-th section of the microstrip and slot, \( l_i \) is a quarter of the effective wavelength for section \( i \) calculated at a centre frequency \( f_c \). The centre frequency for the two sections can be assumed equal at the initial stage of the design. Our simulation results indicate that to get the widest possible bandwidth it is better to design the centre section at about 10% lower than the centre frequency of the assumed band.

**Figure 2** Photo for the developed coupler. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

**Figure 3** Simulated performance of the designed 3 dB coupler. (a) amplitude and (b) phase. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
while the outer sections are to be designed at a frequency which is higher than the centre frequency by about 10%. The relation between $D_3i$ and $l_i$ is given by [7];

$$D_3i = \frac{\sqrt{l_i^2 + \left(0.7855D_1D_2/l_i\right)^2} + l_i}{2} \quad (8)$$

3. RESULTS

The validity of the presented design method is tested by designing and building a 3 dB directional coupler aimed for operation in the 2 to 12 GHz frequency band. For this band, the design frequency is 7 GHz. Rogers RO4003C substrate featuring a dielectric constant of 3.38 and a loss tangent of 0.0027, 0.508 mm thickness plus 17 $\mu$m thick conductive coating is selected for the coupler’s development. Using the proposed design method and with the help of fine tuning using the optimization capability of the software Ansoft HFSSv10, parameters of the coupler are found to be: $D_{11} = 2$ mm, $D_{12} = 6.5$ mm, $D_{21} = 1.8$ mm, $D_{22} = 10.5$ mm, $D_{31} = 6.6$ mm, $D_{32} = 7.4$ mm. It was found that the optimized values of the design parameters are less than 10% different from those obtained by the described design method. This indicates the high accuracy of the method. A photograph of the developed coupler is shown in Figure 2.

The return loss, coupling, isolation and phase behavior of the designed coupler are first verified with the use of CST Microwave Studio, as shown in Figure 3. From the results shown in Figure 3, it is apparent that the designed coupler features desired characteristics across the assumed frequency band. The coupling is $3 \pm 1$ dB for the band 2.1–12 GHz. The isolation is higher than 25 dB and return loss is better than 20 dB (25 dB on average) for the same band. The simulated phase difference between Ports 2 and 3 is $90 \pm 1.5$ degree in the same band.

The developed directional coupler (including SMA ports) is tested using a vector network analyzer. The measured results are shown in Figure 4. The coupler reveals UWB behavior with coupling equal to $3 \pm 1$ dB for the band 2.3 to 12.3 GHz. The isolation and the return loss are better than 23 dB across the band. The measured phase difference between Ports 2 and 3 is $90 \pm 2.5^\circ$ over the same band. These results show good agreement between the measured (see Fig. 4) and simulated (see Fig. 3) performances of the coupler.

4. CONCLUSION

A novel UWB 3dB directional coupler of a compact size accompanied by a simple design method has been described. The structure of the proposed coupler is multilayer with broadside microstrip/slot coupling. The designed coupler shows UWB behavior across the band from 2.3 to 12.3 GHz. Because of its compact size and excellent electrical performance it should be of considerable interest to designers of UWB components.

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Compact Microwave Six-Port Vector Voltmeters for Ultra-Wideband Applications

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Abstract—This paper presents the design of two compact fully integrated six-port devices, which, from scalar power measurements, are capable of determining the ratio of two complex voltages or two complex wave amplitudes over a specified frequency band. Such a task can be performed using a microwave vector network analyzer (VNA), which is one of the most popular microwave measurement instruments in use today. Based on a heterodyning receiver technique, this device is capable of providing a very accurate magnitude and phase measurement of two complex signals over a large dynamic range (90 dB or more). This is possible due to the use of a sophisticated receiver technique that involves a frequency synthesized source and double frequency conversion from a gigahertz region to a 100-kHz region, where signals can be processed using digital techniques. Its excellent performance is offset by its high price tag and a bulky size, which make its use limited to a laboratory environment.

In [1], Hoer and Roe, and in [2] and [3], Engen introduced the concept of a six-port vector voltmeter as a low-cost alternative to the conventional VNA. This device, which is formed by a six-port junction and four scalar power detectors, relies on a low-cost homodyne receiver technique. By performing simple mathematical operations on powers measured at the four chosen ports, it can provide information about the ratio of two complex microwave signals in the remaining two ports. An inclusion of additional couplers and/or power dividers turns this device into a reflectometer or a network analyzer [4]–[6]. For a dynamic range of 60 dB, the power levels at the four detectors vary only within a 15–20-dB range [3], which is convenient from the point of view of signal conditioning for the analog-to-digital conversion. The 60-dB dynamic range can be increased by employing a locking amplifier technique [7]. Due to its small size and much lower manufacturing price, the six-port voltmeter can be used as a portable device in harsh environments. Due to the same reasons, it can be used in multiples, as a receiver connected to individual elements of a large size array antenna.

Our interest in a six-port vector voltmeter is with respect to two application areas, which are microwave imaging [8] and ultra-wideband (UWB) communications [9]. In the first case, we aim to use UWB vector voltmeters in conjunction with tapered slot antennas to form a planar or circular array in a high-resolution microwave camera [8]. With respect to UWB communications, a six-port voltmeter is aimed to be used as a UWB phase detector or modulator/demodulator [9]. In the two cases, our considerations are constrained to the 3.1–10.6-GHz ultra-wide frequency band, as specified in [10].

In the past, wideband six-port voltmeters were assembled using commercially available 90° and 180° couplers and power dividers in stripline, microstrip, and waveguide technology. Since individual couplers and dividers need to be connected, e.g., using coaxial connectors and cables, this approach does not result in an integrated six-port design. Another hurdle is the increased development cost. These commercially available components are not cheap, especially when a UWB performance is of concern. In turn, applying the available designs of UWB couplers and dividers, from the open microwave literature, to form an integrated six-port design is also a challenge. For example, coupled-line 3-dB couplers, which employ the tandem configuration, require crossovers [11]. This creates a challenge to manufacture them even in well-advanced microwave laboratories.

In order to counter this situation, symmetric planar five-ports were studied for some time to obtain low-cost fully integrated six-port voltmeters [12]. Unfortunately, their operational bandwidth is limited to approximately one octave [13]. The largest operational bandwidth of 76% for a planar five-port that employed multiple circular rings and star networks was demonstrated in [14].

The hurdles associated with the design of UWB stripline couplers (caused by the use of crossovers) and the limited operational bandwidth of planar five-ports have triggered our investigations into new configurations of multilayer microstrip couplers and dividers to form a fully integrated UWB six-port vector voltmeter. Our particular interest has focused on complimentary multilayer microstrip and slot structures because of
their potential to enable a UWB performance while avoiding wire crossovers.

In the first step, we choose the multilayer microstrip/slot UWB coupler [16] to form a six-port voltmeter. This coupler, being of broadside coupled lines type, has its ports on two sides of a common ground plane. As a result, it is incompatible with many available UWB power dividers of uniplanar type to realize an integrated six-port voltmeter. An initial solution concerning a multilayer microstrip divider that offers compatibility with the earlier designed UWB microstrip/slot coupler has been described in [15]. In this paper, we provide full details of the solution shown in [15] and point out its shortcomings. As a result of additional research, we propose a new multilayer divider and then we show a new UWB six-port voltmeter, which offers better integration capabilities than its predecessor reported in [15].

II. DESIGN

Following the guidelines presented in [3], a six-port vector voltmeter can be designed using three 3-dB quadrature couplers and one in-phase (0° phase difference between the output ports) or out-of-phase (180° phase difference between the output ports) power divider. In an alternative arrangement, four 3-dB quadrature couplers can be exclusively used to build this device. Here, we concentrate on the design of the vector voltmeter formed by three 3-dB quadrature couplers and one out-of-phase divider, as shown in Fig. 1.

The choice of an elliptical shape for the microstrips and the slot in the conductor supporting the top and bottom dielectric layers. The two layers are coupled via an elliptical slot, which is made in the conductor supporting the top and bottom dielectric layers. The top and bottom conductor layers include elliptically shaped microstrips.

Assuming that the couplers and the divider operate in an ideal manner (as shown by the signal distribution in Fig. 1), the real and imaginary parts of the ratio of two complex signals c and i can be expressed in terms of powers $P_1$ measured by four square-law power detectors as follows:

$$
T = \frac{c}{d} = T_1 + jT_2 = \frac{(P_4 - P_5) + j(P_2 - P_3)}{|\beta|^2}, \quad (1)
$$

This formula can be used for real-time processing or for displaying the ratio of two complex signals. In order to obtain more accurate results for $T$, computer correction techniques involving a suitable calibration stage can be applied, as described in [11–13].

The challenge is to obtain a compact and integrated design of the device shown in Fig. 1, which would operate well, but not necessarily in an ideal manner, over the ultra-wide frequency band of 3.1–10.6 GHz.

In order to obtain a suitable solution to this problem, we apply a multilayer microstrip/slot technology to design a 3-dB quadrature coupler (Q) and an out-of-phase power divider (D). The strategy to design a UWB 3-dB microstrip/slot coupler follows the one that has recently been reported by Abbosh and Bialkowski in [16]. The design of this coupler stems from an extension of the work of de Ronde [17], Garcia [18], and Tanaka et al. [19]. The design shown in [16] outperforms the ones described in [17–19].

The problem is that this coupler cannot be directly integrated with commonly available planar UWB in-phase or out-of-phase power dividers such as the multistage Wilkinson divider. This is because two pairs of its input/output ports appear on two sides of a common ground plane, whereas the Wilkinson divider has its all three ports in one plane.

In order to overcome this problem, in the first instance, we propose to use a UWB out-of-phase divider, which was described in [20]. This power divider employs a parallel strip input port and two microstrip output ports on opposite sides of a ground plane, as required for connecting the microstrip/slot coupler described in [16]. Here, we introduce a taper for the input port of this divider [20] to obtain a better performance.

One shortfall of this power divider is a parallel-strip input port, which prevents its integration with other components having microstrip type ports. In order to overcome this shortfall, we introduce a new design of a UWB out-of-phase divider with three microstrip ports.

The configurations of the 3-dB coupler and two alternative out-of-phase power dividers to form an integrated vector voltmeter are shown in Fig. 2(a)–(c).

The UWB coupler shown in Fig. 2(a) consists of three conductor layers interleaved by two dielectric layers. The top and bottom conductor layers include elliptically shaped microstrips. The two layers are coupled via an elliptical slot, which is made in the conductor supporting the top and bottom dielectric layers.

The choice of an elliptical shape for the microstrips and the slot was found to be advantageous in terms of obtaining high-quality return loss, coupling, and isolation over UWB. The microstrip lines forming the input/output ports of the coupler are designed to have 50-Ω characteristic impedance.

The out-of-phase power dividers shown in Fig. 2(b) and (c) also use two substrates supported by a common ground plane. In the divider shown in Fig. 2(b), the input port is formed by a parallel stripline, which is transformed into two microstrip-line output ports. This arrangement enables equal signal division in magnitude with a 180° difference in phase. The common ground plane is removed in the parallel strip line region; however, it exists in the region of the two microstrip lines. Similarly, as in the coupler of Fig. 2(a), the two microstrip lines are designed to...
have 50-Ω characteristic impedance. Compared with the design described in [20], the configuration shown in Fig. 2(b) uses the Klopfenstein taper in the parallel strip region [15], whereas in [20], the taper has an elliptical shape. There is also an impedance step to compensate for the discontinuity between the microstrip and parallel strip regions.

The out-of-phase power divider shown in Fig. 2(c) has three microstrip ports. The input port can appear on either side of the ground plane, while each of the two output ports is on the opposite side to the other. This allows for the connection of another divider of the same type or a coupler of Fig. 2(a). The divider of Fig. 2(b) employing a parallel strip input port does not provide this flexibility.

The input port (port 1) of the divider in Fig. 2(c) is a tapered microstrip line in order to improve matching between the 50-Ω input port and the junction loaded by two output ports (ports 2 and 3). In addition, this divider uses a UWB microstrip to slot transition and then a UWB T-junction with a vertical slot as a via to the two microstrip output ports. The design of a single microstrip/slot transition follows the ideas described in [21]. Here, this transition is used three times to create a new out-of-phase power divider. Equal signal division in magnitude and a 180° phase difference stems from the symmetry and the fact that the two output microstrip lines run in opposite directions. This 180° out-of-phase signal division was confirmed by a detailed field and signal study using Ansoft’s High Frequency Structure Simulator (HFSS). It is worthwhile to note that if the two output microstrip lines run in the same direction [instead of the opposite directions, as shown in Fig. 2(c)], then the device becomes an in-phase divider with an operational bandwidth similar to that of the out-of-phase divider.

Initial dimensions of the coupler and the two power dividers of Fig. 2 are obtained using the design rules for the coupled lines, parallel strip lines, microstrip lines, and microstrip–slot transitions. For example, for the coupler, we first select the width of the microstrip line \( w_{\text{m}} \) for 50-Ω characteristic impedance. We then apply the even-odd mode analysis for uniform broadside coupled lines [16] to work out the remaining dimensions, which include the length \( l_3 \) of the microstrip/slot and the widths \( w_1 \) and \( w_2 \) of the patch and slot. The length \( l_3 \) governs the frequency of operation, while the widths \( w_1 \) and \( w_2 \) are responsible for the value of the coupling coefficient. They are sensitive design parameters.

Similar design principles are applied to the two dividers. The widths of the microstrip lines \( (w_{\text{m}}) \) and the parallel stripline \((w_{\text{ps}})\) in Fig. 2(b) are chosen for 50-Ω characteristic impedance. The remaining dimensions are chosen using the following guidelines. In the divider of Fig. 2(b), a longer parallel stripline taper \((l_\text{t} + l_\text{s} + l_\text{i})\) is responsible for a better quality return loss. Other parameters, such as line bends, have secondary influence on the performance of this divider.

Similarly, as in its predecessor of Fig. 2(b), the divider of Fig. 2(c) has only a few parameters, which need to be determined. These are the diameter of the virtual open circuit \((r_{\text{oc}})\), the diameter of the virtual short circuit \((r_{\text{sc}})\), the slot width \((s)\), and the slot length. Via computer simulations, we found that the reduction of diameter of the virtual short \((r_{\text{sc}})\) and the open circuit \((r_{\text{oc}})\) leads to shifting the operational band towards higher frequencies. This frequency shift is accompanied by slightly increased insertion losses. This finding is in agreement with the one for a single microstrip/slotline transition [21].

The narrower slot allows for a higher return loss at the input port. However, to avoid manufacturing problems, wider slots are used in our final design. The changes both in the slot length and width have a secondary influence on the divider’s performance. Following these guidelines, we use the formulas described in [20] and [21] to determine the dimensions of this divider.

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**Fig. 2.** Configuration of: (a) UWB 3-dB microstrip/slotline coupler, (b) power divider with a parallel strip input port and two microstrip output ports, and (c) power divider with three microstrip ports.
The initial design of the coupler and the dividers is followed by a fine tuning with full-wave electromagnetic (EM) analysis software (in our case, Ansoft’s HFSS and CST’s Microwave Studio). We found that the design formulas given in [16], [20], and [21] provide a very good approximation to the final dimensions.

III. RESULTS

Two double-sided Rogers RO4003C printed circuit boards (PCBs) were used to develop the prototype couplers, dividers, and then the six-port devices. The chosen substrate features a relative dielectric constant of 3.38 and a loss tangent of 0.0027. It is 0.508-mm thick and includes 17 μm of conductive coating.

In the first step, the coupler of Fig. 2(a) and the dividers of Fig. 2(b) and (c) were designed, manufactured, and tested. Here, we briefly report on the performance of the 3-dB coupler and the first out-of-phase divider [of Fig. 2(b)], whereas more information is given on the out-of-phase divider of Fig. 2(c).

For dimensions of \( l_1 = 4.7 \text{ mm}, \ l_2 = 7.2 \text{ mm}, \ l_3 = 4.8 \text{ mm}, \ \text{and} \ \nu_{\text{out}} = 1.1 \text{ mm}, \) the coupler provides the simulated coupling of 3 dB ± 0.8 dB, the return loss better than 22 dB, and the isolation of not less than 28 dB in the 3.1–10.6-GHz band.

The divider of Fig. 2(b) having dimensions of \( l_1 = 3.21 \text{ mm}, \ l_2 = 1.1 \text{ mm}, \ l_3 = 1.1 \text{ mm}, \ l_4 = 0.25 \text{ mm}, \ l_5 = 1 \text{ mm}, \ l_6 = 2.5 \text{ mm}, \ l_7 = 1 \text{ mm}, \ \nu_1 = 1.7 \text{ mm}, \ \nu_2 = 1.5 \text{ mm}, \ \nu_3 = 1.7 \text{ mm}, \ \text{and} \ l_8 = 7.7 \text{ mm} \) offers the input return loss greater than 16 dB, power division of −3.32 dB ± 0.1 dB, the output return loss greater than 8 dB, and the phase difference between the output ports of 180° ± 1° across the same frequency band. The validity of the simulated results was confirmed by measurements.

The power divider of Fig. 2(c) having dimensions \( l_1 = 1.1 \text{ mm}, \ l_2 = 2 \text{ mm}, \ l_3 = 1 \text{ mm}, \ \text{and} \ s = 0.1 \text{ mm} \) offers the input return loss greater than 13.5 dB, power division of −3.32 dB ± 0.3 dB, the output return loss greater than 8 dB, and phase difference between the output port of 180° ± 0.5° across the same frequency band. This is demonstrated in Fig. 3.

Similarly as for the coupler and first divider, the simulated results were fully confirmed by measurements. The next step was to integrate the individual coupler/dividers into a vector voltmeter.

Outlines of the two fully integrated six-port vector voltmeters (excluding power detectors) are shown in Fig. 4(a) and (b). They were obtained using either Ansoft’s HFSS or CST’s Microwave Studio. The two vector voltometers, designated as vector voltmeter #1 and #2, respectively, differ by the choice of the power divider (D) in the schematic of Fig. 1. The configuration shown in Fig. 4(a) uses the out-of-phase divider of Fig. 2(b), whereas the one in Fig. 4(b) includes the out-of-phase divider of Fig. 2(c).

The two vector voltometers were manufactured and tested. Photographs of the manufactured vector voltmeters (excluding the power detectors) are shown in Fig. 5(a) and (b).

As observed in Fig. 5, the two substrates were affixed using plastic screws to minimize the effect of air gaps. For the testing purposes, the prototypes include subminiature A (SMA) ports.

The overall dimensions of these devices excluding the SMA connectors are only 59 mm × 37 mm for vector voltmeter #1 and 43 mm × 43 mm for vector voltmeter #2. These dimensions confirm the compact size of the developed devices.

In the ideal case, each of the two devices should feature high return losses at ports 1 and 2, high isolation between ports 1 and 2, and 6-dB insertion losses from ports 1 and 2 to ports 4–7. Note that port 3 is reserved for inclusion of a matched load.

Fig. 6 shows both the simulated and measured results for return and insertion losses of vector voltmeter #1.

Note that the measured results include the nonideal performances of the microstrip to SMA transitions and other adverse effects caused by air gaps between the two substrates.

As observed in Fig. 6, the device features simulated return losses greater than 20 dB at port 1 and greater than 25 dB at port 2 across the 3.1–10.6-GHz band. The simulated insertion losses are 6.5 dB ± 1.5 dB over 3.6–10 GHz, which are not far away from the ideal case of 6 dB.

The measured return losses at ports 1 and 2 of vector voltmeter #1 are higher than 15 dB across the 3.1–10.6-GHz band. Insertion losses from port 1 or 2 to the remaining ports are 6.5 dB.
Fig. 4. Outlines of the integrated six-port vector voltmeters employing: (a) divider of Fig. 2(b) and (b) divider of Fig. 2(c).

The measured isolation between ports 1 and 2 (not plotted here) was greater than 20 dB in the specified band of 3.1–10.6 GHz.

The following verification concerns the phase characteristics of the investigated vector voltmeters. When considering the phase of transmission coefficients between port 1 or port 2 and the remaining ports, it is important to check that it is of an appropriate value and stays approximately constant as a function of frequency with respect to a chosen reference port. For the ideal case, illustrated in Fig. 1, they should be integer multiples of 90°. Here we demonstrate that this property is fulfilled for voltmeter #2. The results obtained for voltmeter #1 are similar and, therefore, are not shown here.
Fig. 6. Simulated ($S$) and measured ($S$) results for the $S$-parameters of the six-port vector voltmeter #1. (a) Return losses at ports 1 and 2. (b) Transmission coefficients.

Fig. 7. Simulated ($S$) and measured ($S$) results for the $S$-parameters of the six-port vector voltmeter #2. (a) Return losses at ports 1 and 2. (b) Transmission coefficients.

Fig. 8. Simulated and measured phase characteristics as a function of frequency for vector voltmeter #2. $S_{11}$, $S_{12}$, $S_{21}$, and $S_{22}$ represent the phase of the transmission coefficient $S_{41}$ in Figs. 6(b) and 7(b), with all other phases referenced against $S_{41}$. Subscript m indicates the measured results.

The presented results lead to the conclusion that the developed vector voltmeters exhibit good performance over the aimed...
ultra-wide frequency band from 3.1 to 10.6 GHz. The advantage of voltmeter #2 is its capability to form a fully integrated reflectometer or analyzer [4]–[6], as all of its ports are of microstrip type. Although accomplished in a low dielectric constant substrate (Rogers RO4003), both designs are of a compact size, 59 mm × 37 mm for vector voltmeter #1 and 43 mm × 43 mm for vector voltmeter #2. It is apparent from Fig. 5 that achieving a smaller size is possible by making microstrip ports of shorter length.

The other issue concerns the housing of the manufactured devices. From our full EM simulations, we found that the operation of the individual couplers and of the integrated voltmeters is unaffected by the presence of the enclosure when its height is at least four times greater then the thickness of the double substrate (4 × 2 × 1.5856 mm = 4.151 mm for Rogers RO4003).

IV. CONCLUSION

In this paper, the design of two compact planar vector voltmeters based on a six-port technique, which provide operation over an ultra-wide frequency band from 3.1 to 10.6 GHz, has been presented. The devices are fully integrated and use three couplers and one power divider in microstrip-slot technology. Their initial design has been carried out using an intuitive approach that includes the design rules for coupled lines, parallel strip lines, microstrip lines, and microstrip-slotline transitions. Their fine tuning has been accomplished with the use of commercially available full-wave computer-aided design (CAD) packages. The designed devices have been manufactured and experimentally tested. Their UWB operation has been confirmed both by simulations and measurements. The manufactured devices are ready to be incorporated in high-resolution microwave imaging systems or UWB communication subsystems.

The value of the presented designs is that they use novel solutions to multilayer couplers and dividers whose ports appear on various sides of a common ground plane. The proposed components are compatible so they can be integrated. The presented designs should be of interest to the designers of UWB multi-layer microwave circuits.

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Ultra wideband vertical microstrip–microstrip transition

A.M. Abbosh

Abstract: Simple design guidelines for an ultra wideband aperture-coupled vertical microstrip–microstrip transition are presented. The proposed transition uses broadside coupling between elliptical-shaped microstrip patches at the top and bottom layers via an elliptical-shaped slot in the mid-layer. Theoretical analysis indicates that the best performance concerning the insertion loss and the return loss over the maximum possible bandwidth can be achieved when the distance between the top and bottom coupled patches is equal to 0.8 (or 1.94 dB). Simulated and measured results show that the proposed transition has an insertion loss of <0.7 dB and a return loss of >15 dB across the frequency band 3.1–10.6 GHz.

1 Introduction

Vertical microstrip–microstrip transitions are essential to build compact microwave circuits in multilayer designs such as monolithic microwave-integrated circuits and low-temperature co-fired ceramic circuits (LTCC). In these applications, the transitions are usually designed to have a minimum insertion loss over the maximum possible bandwidth.

Vertical transitions have been extensively studied. A literature review of the vertical transitions shows that they are either aperture-coupled transitions or via-hole transitions [1–16]. As a general rule, it was noticed that as the operating frequency increases, the performance of the via-hole transitions is degraded. Their insertion loss increases and bandwidth is deteriorated. In addition, their fabrication process is difficult. Earlier design of the via-hole transitions used a single via-hole to connect transmission lines that are thin enough to fit the diameter of one via-hole [1–6]. When the substrate used is thick or has a low permittivity, the transmission line becomes too wide to be covered with only one via-hole. To solve this problem, a vertical microstrip-to-microstrip metallic wall was used to build a broadband transition [7]. Owing to the difficulty in fabricating the metallic wall introduced in [7] was implemented in the form of an array of several via-holes previously proposed in [8]. The results shown in [7] reveal that as the operating frequency increases (>6 GHz), the insertion loss becomes relatively high (>1 dB).

To overcome the shortcomings of the via-holes, aperture-coupled vertical transitions are normally used, although their relatively high insertion loss problem is still to be solved. In [9], an aperture-coupled transition structure was made with an electrically wide aperture formed on the common ground plane of a two-layered structure. The proposed structure was verified using the method of moment analysis and measurements. The measured results show a high insertion loss (~2.7 dB).

A literature review indicates that there are some other configurations of the vertical aperture-coupled transitions proposed for specific applications. Lin [10] presented a vertical aperture-coupled transition between a microstrip line and a coplanar waveguide and limited to the C-band, whereas Lafond et al. [11] designed a vertical transition for special millimetre-range applications, where very thin substrates are supported by a thick ground plane. In the recent article [12], a vertical transition between coplanar waveguides was used to build band pass filters. The used transition has narrow band behaviour and cannot be considered for broadband applications.

Recently, a design for broadband vertical transitions using microstrip-fed cavity couplers was presented [13–15]. The design features a cavity in an electrically thick ground plane between two parallel back-to-back microstrip lines terminated by open-circuited stubs. The analysis and design of that cavity was originally presented by Pozar [16] for couplers, dividers and vertical transitions. The difference between the cavity used in [13] and the one used in [14, 15] is that it is made with vias and filled with dielectric material in the former, whereas in the latter, an air cavity is made with perfect metallic walls. An additional metallic cavity layer with thickness equal to or greater than that of the dielectric layer is required between the printed circuit board (PCB) signal layers. This additional layer increases the complexity of the design and makes it incompatible with thin LTCC technology. Moreover, the results shown in [13, 14] indicated narrow band characteristics in the millimetre range with ≥2 dB insertion loss, whereas the transition in [15] can only cover a part of the ultra wideband (UWB) (3.1–10.6 GHz) with an insertion loss of ≥1 dB.

In this article, simple design guidelines are proposed for an aperture-coupled vertical microstrip–microstrip transition. In this design, an elliptical structure is considered for the microstrip-coupled patches and the coupling slot. It is to be noted that the elliptical shape for the coupled structure is chosen because of its ability to achieve an almost constant coupling value over the UWB. This ability comes from the fact that the elliptical shape resembles a tapered coupled structure. The simulated and measured results show that the proposed transition has a
low insertion loss (<0.7 dB) and a high return loss (>15 dB) over the UWB (3.1–10.6 GHz).

2 Design

The Configuration of the proposed transition is shown in Fig. 1. It consists of two elliptical microstrip patches that are connected to the input and output microstrip lines and facing each other at the top and bottom layers. The coupling between these patches is achieved via an elliptical slot in the ground plane, which is located at the mid-layer. The proposed structure is designed using the odd–even analysis of coupled microstrip lines [17–20].

The analysis starts by considering the transition as a four-port backward coupler with two of the output ports terminated in an open circuit (Fig. 2). Assume that the device is designed to have a coupling equal to \( C \) between the top and bottom patches and that the input and output signals to/from the \( n \)th port are \( a_i \) and \( b_i \), respectively. Depending on principles of the four-port backward coupler [17–19], the output signals at the input and output ports can be calculated as follows

\[
\begin{align*}
  b_1 &= \frac{jC \sin(\beta_{ef} l) a_3 + \sqrt{1 - C^2} a_4}{\sqrt{1 - C^2 \cos(\beta_{ef} l) + j \sin(\beta_{ef} l)}} \\
  b_2 &= \frac{jC \sin(\beta_{ef} l) a_4 + \sqrt{1 - C^2} a_3}{\sqrt{1 - C^2 \cos(\beta_{ef} l) + j \sin(\beta_{ef} l)}}
\end{align*}
\]

(1) and (2)

where \( l \) is the length of the coupling structure and \( \beta_{ef} \) is the effective phase constant in the medium of the coupling structure. For the structure under investigation, it is possible to show that

\[
\beta_{ef} = \beta_e + \beta_o = \frac{2 \pi \sqrt{\varepsilon_r}}{\lambda}
\]

(3)

where \( \beta_e \) and \( \beta_o \) are the phase constants for the even and odd modes, respectively, \( \lambda \) the free space wavelength and \( \varepsilon_r \) the dielectric constant of the substrate.

As ports 3 and 4 are terminated in an open circuit, the reflection coefficient at those ports is equal to 1 and assuming that the output port 2 is perfectly matched, the incident (i.e., reflected) signals at ports 3 and 4 are

\[
\begin{align*}
  b_3 &= \frac{jC \sin(\beta_{ef} l) a_1}{\sqrt{1 - C^2 \cos(\beta_{ef} l) + j \sin(\beta_{ef} l)}} \\
  b_4 &= \frac{\sqrt{1 - C^2} a_1}{\sqrt{1 - C^2 \cos(\beta_{ef} l) + j \sin(\beta_{ef} l)}}
\end{align*}
\]

(4) and (5)

From (1–5) and knowing that \( S_{11} = b_1/a_1, S_{21} = b_2/a_1 \) are the return loss of the input port and the insertion loss from the input port to the output port, respectively, then

\[
\begin{align*}
  S_{11} &= \frac{1 - C^2(1 + \sin^2(\beta_{ef} l))}{\sqrt{1 - C^2 \cos(\beta_{ef} l) + j \sin(\beta_{ef} l))^2}} \\
  S_{21} &= \frac{j2C \sqrt{1 - C^2 \sin(\beta_{ef} l)}}{\sqrt{1 - C^2 \cos(\beta_{ef} l) + j \sin(\beta_{ef} l))^2}}
\end{align*}
\]

(6) and (7)

It is to be noted that (6) and (7) are related to each other via \(|S_{11}|^2 = 1 - |S_{21}|^2\) as the coupled structure is assumed to be reciprocal lossless.

Designing a high performance transition requires that \( S_{11} = 0 \) (or \(-\infty\) dB) and \( S_{21} = 1 \) (or 0 dB) over the maximum possible bandwidth and, in this paper, the objective is the UWB 3.1–10.6 GHz. To achieve these requirements, it is possible either to choose a certain value for \( C \) and calculate the optimum effective electrical length of the coupled region (\( \beta_{ef} l \)) from (6) (or (7)), or vice versa. In Figs 3 and 4, variations of the insertion loss and return loss with the effective coupling length \( \beta_{ef} l \) are shown for different values of coupling \( C \). It is clear that in order to achieve the best performance (low insertion loss and high return loss) over a broadband, then \( \beta_{ef} l \) should be equal to 90° at the centre frequency of operation. Concerning the optimum value for \( C \), it is obvious from Fig. 3 that the widest 10 dB return loss bandwidth occurs when \( C = 0.8 \) (or 1.94 dB) with an acceptable insertion loss, which is <0.4 dB, as shown in Fig. 4. It is to be noted from Figs 3 and 4 that the lowest insertion loss at the centre frequency occurs when \( C = 0.7 \) (or 3 dB), but this value gives a lower bandwidth when compared with the case \( C = 0.8 \).

Fig. 1 Configuration of the proposed transition

(a) Top layer
(b) Mid-layer
(c) Bottom layer
(d) Whole structure
Before designing the transition, it is important to make sure that it covers the UWB 3.1–10.6 GHz with acceptable performance, for example, insertion loss <0.5 dB and return loss >10 dB. As mentioned earlier, the coupling length of the transition is equal to 90° at the centre frequency, which is (3.1 + 10.6)/2 = 6.85 GHz. Hence, the coupling length is equivalent to 90 × 3.1/6.85 = 40.7° at 3.1 GHz and 90 × 10.6/6.85 = 139.3° at 10.6 GHz. According to Figs. 3 and 4, the return loss and the insertion loss at C = 0.8 are equal to 17 and 0.1 dB, respectively. These predicted values prove that the transition at C = 0.8 is the optimum choice for UWB applications. In order to make a fair comparison between the theoretical estimation of the performance and the measured and simulated results, the two transitions are to be built and tested.

Depending on the value of the coupling, the even (Z_{oe}) and odd (Z_{oo}) mode characteristic impedances for the coupling patches are calculated using the following equation

\[
Z_{oe} = Z_o \sqrt{\frac{1 + C}{1 - C}}, \quad Z_{oo} = Z_o \sqrt{\frac{1 - C}{1 + C}}
\]

where \(Z_o (=50 \Omega)\) is the characteristic impedance of the microstrip ports of the coupler. Using C = 0.8 for the widest bandwidth, the impedances \(Z_{oe}\) and \(Z_{oo}\) are found to be 150 and 16.7 \(\Omega\), respectively. If the transition is designed to have the least possible insertion loss at the centre frequency, then C = 0.7; hence \(Z_{oe}\) and \(Z_{oo}\) are found to be 120.7 and 20.7 \(\Omega\), respectively. To calculate the transition dimension which gives these impedance values, it is possible to use the following equation [20]

\[
Z_{oe} = \frac{60\pi K(k_1)}{\sqrt{\varepsilon_r K(k_1)}}, \quad Z_{oo} = \frac{60\pi K'(k_2)}{\sqrt{\varepsilon_r K(k_2)}}
\]

where \(\varepsilon_r\) is the dielectric constant of the substrate, \(K(k)\) the first kind elliptical integral and \(K'(k) = K(\sqrt{1 - k^2})\).

The above-mentioned method was applied to design and build a vertical microstrip–microstrip aperture-coupled transition operating over the UWB frequency range from 3.1 to 10.6 GHz. Rogers RO4003C with thickness 0.508 mm, dielectric constant 3.38 and tangent loss 0.0023 was used as a substrate. The design process was aided with a full electromagnetic simulation package (Ansoft HFSSv10), whereas the measurements were accomplished using a vector network analyser in an anechoic chamber.

Using the proposed design method and with the help of fine tuning using the optimisation capability of the software Ansoft HFSSv10, parameters of the transition with \(C = 0.8\) were found to be 5.4, 7.9, 7.2 and 1.2 mm for \(D_m, D_s, D_{sec}\) and \(w_m\), respectively. For the transition with \(C = 0.7\), the...
dimensions are 4.8, 7.4, 7.2 and 1.2 mm for $D_m$, $D_e$, $D_{sec}$ and $w_m$, respectively. It was found that the optimised values of the design parameters are <10% different from those obtained by the design method described. This indicates the high accuracy of the method. A photograph of one of the developed transitions is shown in Fig. 5. It was built on a substrate with an overall dimension of $2 \times 3 \text{ cm}^2$, including the input/output microstrip transmission lines.

Fig. 6 shows the simulated and measured return losses of the input/output ports of the transitions. The average simulated and measured return loss is $\approx 15 \text{ dB}$, when $C = 0.8$, across the whole UWB. Concerning the $C = 0.7$ transition, the return loss at the centre frequency is better than 30 dB, whereas it is 7 dB at the low and high end of the UWB. The results shown in Fig. 6 are in good agreement with those of the theoretical analysis presented in Fig. 3. A small discrepancy is due to the fact that the results shown in Fig. 3 assume a constant value for coupling $C$, whereas in reality, the coupling varies slightly across the UWB, especially at the two ends as shown in [21].

The simulated and measured insertion losses of the developed transitions are presented in Fig. 7. The results indicate a UWB behaviour over the band 3.1–10.6 GHz, with an insertion loss that is $<0.6 \text{ dB}$ according to the simulation and 0.7 dB according to the measurement for the transition with $C = 0.8$. For the transition $C = 0.7$, the insertion loss at the centre frequency is as low as 0.2 dB, whereas it is $\approx 1.2 \text{ dB}$ at the two ends of the UWB. The general behaviour of the results presented in Fig. 7 is in good agreement with the theoretical results shown in Fig. 4 and once again a small discrepancy is expected because of a non-constant value of the coupling $C$ across the whole UWB. It is to be noted that the performance of the transition proposed in this article is better than the recently developed aperture-coupled transition [15] using even a simpler structure with clear design guidelines.

There is a small discrepancy between the simulated and measured results shown in Figs. 6 and 7. The manual alignment of the two layers forming the transition and the effect of the subminiature A (SMA) connectors, which were used in the measurements, but not included in the simulation, are the main reasons for this difference.

4 Conclusion

Simple design guidelines for a UWB aperture-coupled vertical microstrip–microstrip transition have been presented. The proposed transition uses broadside coupling between elliptical-shaped microstrip patches at the top and bottom layers via an elliptical-shaped slot in the mid-layer. Theoretical analysis indicated that the best performance concerning the insertion loss and the return loss over the maximum possible bandwidth can be achieved when the coupling between the top and bottom patches is equal to 0.8 (or 1.94 dB). Simulated and measured results have shown that the proposed transition has an insertion loss of $<0.7 \text{ dB}$ and a return loss $>15 \text{ dB}$ across the frequency band $3.1–10.6 \text{ GHz}$.

5 References

ULTRAWIDEBAND APERTURE-COUPLED VERTICAL MICROSTRIP TRANSITION

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ABSTRACT: Simple design guidelines for an ultrawideband aperture-coupled vertical microstrip-microstrip transition are presented. The proposed transition utilizes broadside coupling between elliptical-shaped microstrip patches at the top and bottom layers via an elliptical-shaped slot in the ground plane, which is located at the mid layer. The simulated and measured results show that the proposed transition has a low insertion loss (less than 0.75 dB) and a high return loss (more than 13 dB) over the frequency band 3.1–10.6 GHz. Moreover, the new design cannot cover the operating frequency increases above 6 GHz. The design presented in [5] shows the fabrication challenges and difficulties when trying to improve the performance of the via-holes. The results presented in the paper shows that the performance degrades, i.e. the insertion loss becomes relatively high (more than 1 dB), as the operating frequency increases (above 6 GHz). To overcome these problems, slot-coupled vertical transitions can be used although their relatively high insertion loss problem is still to be solved. In [3] a slot-coupled transition structure was made with an electrically wide aperture formed on the common ground plane of a two-layered structure. The measured results show a high insertion loss (~2.7 dB).

A design for broadband vertical transitions using microstrip-fed cavity couplers was recently presented [4]. The design features a metallic cavity layer in an electrically thick ground plane between two parallel back-to-back microstrip lines terminated by open-circuited stubs. The use of the additional cavity layer adds a complexity to the design and makes it incompatible with thin LTCC technology. Moreover, the new design cannot cover the UWB (3.1–10.6 GHz), and it has an insertion loss around 1 dB across the bandwidth 3–7.5 GHz.

In this paper simple design guidelines are presented for a slot-coupled vertical microstrip–microstrip transition. In this design an elliptical structure is considered for the microstrip coupling patches and the coupling slot. The simulated and measured results show that the proposed transition has a low insertion loss (less than 0.75 dB) and a high return loss (more than 13 dB) over the ultrawideband 3.1–10.6 GHz.

1. INTRODUCTION

Vertical microstrip–microstrip transitions are essential to build compact microwave circuits in multilayer designs such as monolithic microwave integrated circuits and low-temperature cofired ceramic circuits (LTCC). They are either aperture coupled or via-hole transitions [1–5]. All types of transitions are usually designed to have minimum insertion loss over a maximum possible bandwidth.

Concerning the via-hole transitions, it was noticed that as the operating frequency increases the performance of the via-holes is degraded. Their insertion loss increases and bandwidth is deteriorated [4]. The design presented in [5] shows the fabrication challenges and difficulties when trying to improve the performance of the via-holes. The results presented in the paper shows that the performance degrades, i.e. the insertion loss becomes relatively high (more than 1 dB), as the operating frequency increases (above 6 GHz).

To overcome these problems, slot-coupled vertical transitions can be used although their relatively high insertion loss problem is still to be solved. In [3] a slot-coupled transition structure was made with an electrically wide aperture formed on the common ground plane of a two-layered structure. The measured results show a high insertion loss (~2.7 dB).

A design for broadband vertical transitions using microstrip-fed cavity couplers was recently presented [4]. The design features a metallic cavity layer in an electrically thick ground plane between two parallel back-to-back microstrip lines terminated by open-circuited stubs. The use of the additional cavity layer adds a complexity to the design and makes it incompatible with thin LTCC technology. Moreover, the new design cannot cover the UWB (3.1–10.6 GHz), and it has an insertion loss around 1 dB across the bandwidth 3–7.5 GHz.

2. DESIGN

Configuration of the proposed transition is shown in Figure 1. It consists of two elliptical microstrip patches that are connected to the input and output microstrip lines and facing each other at the top and bottom layer. The broadside coupling between these patches is achieved via an elliptical slot in the ground plane, which is located at the mid layer.

The design of the proposed structure is done using the odd–even analysis of broadside coupled microstrip lines [6]. Assume that it is required to have C coupling between the top and bottom patches. The even \( z_{0e} \) and odd \( z_{0o} \) mode characteristic impedances for the coupling patches are calculated using the following equations:

\[
Z_{0e} = Z_0 - \frac{1 + C}{1 - C} Z_{0o} = Z_0 - \frac{1 - C}{1 + C}
\]  
(1)

where \( Z_0 \) (= 50 Ω) is the characteristic impedance of the microstrip ports of the coupler. To achieve a perfect coupling between the top and bottom layers then \( C = 1 \). Hence, from Eq. (1), \( z_{0e} \to \infty \) and \( z_{0o} \to 0 \). To calculate the transition dimension which gives these impedance values, it is possible to use the following equations:

\[
Z_{0o} = \frac{60 \pi K(k_x)}{\sqrt{\varepsilon_r K^*(k_x)}}; Z_{0e} = \frac{60 \pi K'(k_x)}{\sqrt{\varepsilon_r K(k_x)}}
\]  
(2)

where \( \varepsilon_r \) is the dielectric constant of the substrate, \( K(k) \) is the first kind elliptical integral, and \( K^*(k) = K(\sqrt{1-k^2}) \). The parameters \( k_x \) and \( k_x \) are related to dimension of the coupling structure as given in [7].

To get an ideal transition then from Eqs. (1) and (2) and Ref. 7 it is possible to show that dimensions of the coupling structure should be chosen according to the following equation:

\[
D_x/h >> 1; D_y/D_x >> 1; D_y >> D_m >> h
\]  
(3)

Figure 1 Configuration of the proposed transition. (a) Top layer, (b) mid layer, (c) bottom layer, and (d) the whole structure. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
where \(D_m\) and \(D_s\) are the major diameters of the coupling microstrip and slot, respectively, and \(h\) is the substrate thickness.

According to Eq. (3), dimensions of the elliptical coupling structure should be large compared to the substrate thickness. To build a compact transition two options are available; the first is to build the transition using thin-film technology, where \(h\) is equal to tens of microns. Unfortunately this option is not available to the author. The second option is to use an ordinary PCB substrate. In this case and according to Eq. (3), dimensions of the microstrip coupling patches at the top and bottom layers as well as the slot are large compared with \(h\) and the wavelength. It is expected that the radiation losses become high, as the top and bottom coupling structures behave like patch antennas, degrading the performance of the transition and causing undesired interference between different components of the RF front end.

To make a fair compromise between the requirement of increasing the dimensions (to increase the coupling and hence decrease the insertion loss) and decreasing the dimensions (to decrease the radiation losses) it is possible to design a transition that has a certain controlled and acceptable insertion loss, say 0.5 dB. In this case, \(C = 10^{-0.5/20} = 0.944\). Because of the availability of ordinary substrates the author chose the second option.

Solving Eq. (1) for \(C = 0.994\) gives \(z_{0m} = 294.6\ \Omega\) and \(z_{0u} = 8.5\ \Omega\). Major diameters of the elliptical coupling structure \((D_m\) and \(D_s\)) can be found using Eqs. (2), (3) and the method given in [7]. The secondary diameter of the elliptical microstrip/slot coupling structures \((D_{sec})\) is assumed to be equal to quarter of the effective wavelength at the centre frequency of operation, i.e. at 6.85 GHz.

The simulated and measured insertion loss of the developed transition is shown in Figure 3. The results indicate ultrawideband behavior over the band 3.1–10.6 GHz with insertion loss which is less than 0.6 dB in the simulation and 0.75 dB according to the measured results. This is close to the designed value which is 0.5 dB and better than the recently developed aperture coupled transition [4] using even a simpler structure with clear design guidelines.

Figure 4 shows the simulated and measured return loss of the input/output ports of the transition. Note that because of symmetry, return loss of the input port \(S_{11}\) is equal to return loss of the output port \(S_{22}\) and thus \(S_{22}\) is not shown explicitly. The average simulated and measured return loss is around 20 dB across the whole ultrawideband. The simulated return loss is always higher than 15 dB whereas the measured return loss is better than 13 dB over the band 3.1–10.6 GHz.

The last step of the design is to find the width of microstrip transmission lines at the top and bottom layers \((w_m)\) to give 50 \(\Omega\) impedance. This can be achieved using the standard microstrip design equations [8].

### 3. RESULTS AND DISCUSSION

The above outlined method was applied to design and build a vertical aperture-coupled microstrip–microstrip transition operating over the UWB frequency range from 3.1 to 10.6 GHz. Rogers RO4003C with thickness 0.508 mm, dielectric constant 3.38, and tangent loss 0.0023 was used as a substrate. The design process was aided with a full electromagnetic simulation package (Ansoft HFSSv10) while the measurements were accomplished using a vector network analyzer in an anechoic chamber.

Using the proposed design method and with the help of fine tuning using the optimization capability of the software, parameters of the transition were found to be: 0.75 cm, 1.74 cm, 0.67 cm, and 1.2 mm for \(D_m\), \(D_s\), \(D_{sec}\), and \(w_m\) respectively. It was found that the optimized values of the design parameters are around 8% different from those obtained by the described design method. This indicates the acceptable accuracy of the proposed method. A photograph of the developed transition is shown in Figure 2. It was built on a PCB with overall dimension of 3 cm \(\times\) 4 cm.

The simulated and measured return loss of the developed transition is shown in Figure 3. The results indicate ultrawideband behavior over the band 3.1–10.6 GHz with insertion loss which is less than 0.6 dB in the simulation and 0.75 dB according to the measured results. This is close to the designed value which is 0.5 dB and better than the recently developed aperture coupled transition [4] using even a simpler structure with clear design guidelines.

Figure 4 shows the simulated and measured return loss of the input/output ports of the transition. Note that because of symmetry, return loss of the input port \(S_{11}\) is equal to return loss of the output port \(S_{22}\) and thus \(S_{22}\) is not shown explicitly. The average simulated and measured return loss is around 20 dB across the whole ultrawideband. The simulated return loss is always higher than 15 dB whereas the measured return loss is better than 13 dB over the band 3.1–10.6 GHz.
The simulated and measured results presented in Figures 3 and 4 are in good agreement. There is a little discrepancy between them. The manual alignment of the two layers forming the transition and effect of the SMA connectors which were used in the measurements, but not included in the simulation, are the main reasons for this discrepancy.

4. CONCLUSION

Simple design guidelines for an ultrawideband aperture-coupled vertical microstrip–microstrip transition have been presented. The proposed transition utilizes broadside coupling between elliptical-shaped microstrip patches at the top and bottom layers via an elliptical-shaped slot in the ground plane, which is located at the mid layer. The simulated and measured results have shown that the proposed transition has less than 0.75 dB insertion loss and more than 13 dB return loss across the frequency band 3.1–10.6 GHz.

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BROADBAND QUADRATURE COUPLER WITH SLOTTED GROUND PLANE

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ABSTRACT: The design of an edge-coupled quadrature directional coupler, which has a broadband performance and relaxed coupled-line spacing, is presented. A slotted ground plane is used underneath the coupled region in order to relax the requirement for a narrow slot between the coupled lines. The simulated and measured results show that the designed coupler exhibits a coupling of 3 ± 1 dB across the band 3.8 GHz to 9.8 GHz, where the spacing between the coupled lines is 0.12 mm. Without the slot in the ground plane, the spacing should be less than 0.001 mm to achieve the same value of coupling across that band. This broadband coupling is accompanied by an isolation and return loss in the order of 13 dB or better across the band 3–10 GHz. The designed device has a compact size with a dimension of 30 mm × 20 mm. © 2007 Wiley Periodicals, Inc. Microwave Opt Technol Lett 50: 328 – 331, 2008; Published online in Wiley InterScience (www.interscience.wiley.com).
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Key words: directional coupler; broadband; coupled lines

1. INTRODUCTION

Directional couplers using planar transverse electromagnetic lines such as coupled striplines were developed in the mid-1950s [1, 2]. Numerous articles were published after that describing the theory, design, and fabrication of the edge-coupled directional couplers. Those articles cover the loose coupling range, i.e., more than 10 dB. In several applications, a tight coupler with a 3 dB coupling is required. Using an edge-coupled structure, it is difficult to realize this value of coupling as a very narrow spacing is required, which makes the manufacturing process difficult. Several methods have been used to alleviate this problem. In [3], a design which uses a three-dielectric layer broadside-coupled striplines is presented, whereas a tandem connection was suggested in [4]. The tapered parallel-coupled lines were presented in [5] as a possible solution to the applications that require a tight coupling.

In addition to the problem of the narrow spacing required to achieve a tight coupling, another problem faced the design of microstrip directional couplers. Because the microstrip line is inhomogeneous, the even- and odd-mode propagation velocities for a coupled pair of microstrip lines are not equal resulting in a poor directivity. Several techniques were proposed as a solution to this problem, such as the wiggles lines [6], dielectric overlay [7], capacitive compensation [8], and the interdigital structure [9].

To avoid the narrow spacing requirement of the edge-coupled directional couplers, Abbosh and Bielkowski [10] proposed the use of a microstrip-slot-microstrip broadside-coupled structure. They built quadrature couplers which have ultra-wideband performance. The two outputs of the designed couplers are not at the same layer. This could not be suitable for some uniplanar applications. In another development, composite left-right handed coplanar waveguides were used to design a broadband quadrature directional coupler [11]. The use of multiple crossover shorting strips complicates the design. Moreover, the results show a high amplitude imbalance between the two outputs.

In this article, a slotted ground plane underneath the coupled microstrip lines is used to alleviate the need for a narrow spacing between the coupled lines. A complete design method based on the even- and odd-mode analysis with the help of the conformal mapping and image techniques is presented. The method is used to design and build a broadband 3 dB directional coupler, which has a compact dimension with a coupled-line spacing that is easy to manufacture.

2. DESIGN METHOD

The configuration of the proposed directional coupler is shown in Figure 1. The top layer contains the two coupled microstrip lines, whereas the ground plane is located at the bottom layer. There is a rectangular slot made at the ground plane underneath the coupled lines.

The structure presented in Figure 1 can be fully analyzed using the even- and odd-mode of operation. Distribution of the electric field lines between the coupled lines is shown in Figure 2. According to the characteristics of the backward directional couplers, the input port (port1) and port4 are isolated, see Figure 1. Therefore, the analysis that follows concentrates only on the calculation of the coupling between the input port and the coupled output (port2). The power output from port3 can be calculated depending on the value of the input power and the coupled output power.

Figure 1 Configuration of the proposed directional coupler. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
where even- and odd-mode for line mutual capacitance per unit length between the lines.

Even- and odd-mode excitation for the coupled lines. Figure 2 shows the two modes can be found using the relation \[ C_{i0} = e_{0}K(k_{0}^{2})K'(k_{0}) \]

where \( w_{s} \) is the width of each of the coupled lines, \( s \) is the spacing between the coupled lines, \( w_{g} \) is the width of the slot at the ground plane, \( e_{0} \) is permittivity of the air, \( h \) and \( e_{r} \) are thickness and dielectric constant of the substrate, respectively. \( K(k) \) is the first kind elliptical integral and \( K'(k) = K(\sqrt{1 - k^{2}}) \). The parameters \( q \), \( k_{e} \), and \( k_{o} \) in (4) and (5) are functions of the coupler dimension as shown in the following equations, which are the result of the conformal mapping and image techniques,

\[
q = 1 + 1.12(\sqrt{b' + 0.25(w_{s} - s)^{2}/w_{c}})^{0.82}
\]

\[
k_{i} = \frac{s}{s + 2w_{s}}; k_{o} = \frac{\tanh(\pi s/(2h))}{\tanh(\pi(s + 2w_{s})/(4h))}
\]

Assuming that the coupler is required to have \( C \) coupling factor, the even \( (Z_{e}) \) and odd \( (Z_{o}) \) mode characteristic impedances to achieve this value of the coupling are calculated as follows:

\[
Z_{e} = Z_{o}\left(1 + C\right)^{0.5}; Z_{o} = Z_{e}\left(1 - C\right)^{0.5}
\]

where \( Z_{o} \) is the characteristic impedance of the microstrip ports of the coupler.

If \( Z_{e} = 50 \) \( \Omega \) and the coupling factor \( C \) is 0.707 (or 3 dB) then values of \( Z_{o} \) and \( Z_{e} \) can be calculated from (8) and are given as follows: \( Z_{e} = 120.5 \) \( \Omega \) and \( Z_{o} = 20.7 \) \( \Omega \). Using (1)–(7), it is possible find the required dimension for the coupled lines.

The last step of the design procedure concerns the determination of the coupler’s length \( l \). Here, \( l \) was chosen to be quarter of the effective wavelength for the microstrip line at the centre frequency of operation. It can be calculated using the standard microstrip formulas [12].

**Figure 2** Even- and odd-mode excitation for the coupled lines. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
3. RESULTS

The validity of the presented design method was tested by designing and manufacturing a 3 dB directional coupler aimed for the operation within the ultrawideband range (3.1–10.6 GHz). For this band, the center frequency of operation is 6.85 GHz. Rogers RO4003C (with \( \varepsilon_r = 3.38, h = 0.508 \text{ mm}, \text{ loss tangent} = 0.0027 \)) was selected as a substrate for the coupler development.

Using the proposed method with the help of the optimization capability of the software HFSSv10, dimensions of the designed directional coupler (\( w, w_c, l \), and \( s \)) are equal to 0.95 mm, 6.8 mm, 6.95 mm, and 0.12 mm, respectively. It is to be noted that the calculated values for the design parameters (\( w_c, w, l \), and \( s \)) are 0.88 mm, 7.2 mm, 6.8 mm, and 0.12 mm, respectively, indicating the accuracy of the design method where the difference between the calculated and the optimized values is less than 8%. It is worthy to mention that without using the proposed slotted ground plane, the spacing between the coupled lines should be less than 0.001 mm to achieve the required 3 dB coupling. This value for spacing is impractical assuming the use of the general milling and chemical lithography machines as any error due to the manufacturing process degrades the performance to a large extent. The use of the slotted ground proposed in this article relaxed the spacing requirement between the coupled lines by more than 100 times.

The return loss, coupling and isolation of the designed couplers were first verified using the software Ansoft HFSSv10 and then measured using a vector network analyzer. Figure 4 shows the simulated and measured amplitudes of the scattering parameters for the designed 3 dB coupler. These are followed by the results of the phase difference between the two output ports and the group delay from the input to the output ports as shown in Figure 5. It is clear that the designed coupler features broadband characteristics. The coupling is \( 3 \pm 1 \text{ dB} \) across the band 3.8–9.8 GHz. The isolation and the return loss are better than 13 dB across the band 3–10 GHz.

Concerning the phase performance of the manufactured coupler as depicted in Figure 5, it is observed that the phase difference between the output ports (ports 2 and 3) is \( 90^\circ \pm 5^\circ \) over the band 3–9.8 GHz. This result reveals a broadband performance with high phase stability.

In addition to the amplitude and phase balances, the demand of a constant group delay is very crucial for the wideband couplers. Figure 5 shows the variation of the insertion loss, return loss, and isolation of the directional coupler with frequency. (Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com) The simulated group delay for the two output ports are shown in Figure 5. The results indicate an almost constant group delay with very low standard deviations, which are 0.0043 ns for the coupled port and 0.003 ns for the direct port.

4. CONCLUSION

The design of an edge-coupled quadrature directional coupler, which has a broadband performance and relaxed coupled-line spacing, has been presented. A slotted ground plane was used underneath the coupled region to relax the requirement for a narrow slot between the coupled lines. The simulated results have shown that the designed coupler exhibits a coupling of \( 3 \pm 1 \text{ dB} \) across the band 3.8–9.8 GHz, when the spacing between the coupled lines is 0.12 mm. Without the slot in the ground plane, the spacing should be less than 0.001 mm to achieve the same value of coupling across that band. This broadband coupling is accompanied by an isolation and return loss in the order of 13 dB across the band 3–10 GHz. The designed device has a compact size with a dimension of 30 mm \( \times \) 20 mm.

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ANALYTICAL PERFORMANCE EVALUATION OF AlGaN/GaN METAL INSULATOR SEMICONDUCTOR HETEROSTRUCTURE FIELD EFFECT TRANSISTOR AND ITS COMPARISON WITH CONVENTIONAL HFETs FOR HIGH POWER MICROWAVE APPLICATIONS

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ABSTRACT: In this work, a comprehensive analytical model for AlGaN/GaN MISHFET has been presented to evaluate the drain current characteristics, transconductance, and cut-off frequency of the insulated device. The model takes into account the fundamental dependence of the carrier density on position of quasi Fermi level to consider the quantum effects and validate it from subthreshold to high conduction region. The effect of spontaneous and piezoelectric polarization at the AlGaN/GaN interface and parasitic source/drain resistances have also been incorporated in the analysis. Its advantages over conventional HFET structure are discussed in detail. For a MISHFET with quarter micron gate length, the cut-off frequency is reported to be 52 GHz. The MISHFET shows remarkable 36% increase in drain saturation current. The model has a broad utility as it is equally applicable to HFETs as well. The present model is based on closed form expression and does not involve any fitting parameter. The results obtained are compared with experimental data and show excellent agreement, thereby proving the validity of the model. © 2007 Wiley Periodicals, Inc. Microwave Opt Technol Lett 50: 331–338, 2008; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.23073

Key words: AlGaN/GaN MISHFET; polarization; sheet carrier density; gate voltage swing; saturation drain current

1. INTRODUCTION

The wide band gap AlGaN/GaN high electron mobility transistors (HEMTs) show great promise for applications such as high frequency wireless base stations and broad-band links, commercial and military radar and satellite communications [1-5]. The outstanding properties of nitride material system such as high electron mobility, high saturation velocity, low thermal impedance, and high breakdown field make them extremely promising devices for high power and high temperature microwave applications. GaN-based materials are usually grown in [0001] and [111] directions, and since these axes are polar, they cause GaN-based materials to exhibit strong lattice polarization effects. Because of the piezoelectric and spontaneous polarization fields, AlGaN/GaN-based HFETs have the ability to achieve two dimensional electron gas (2-DEG) with sheet carrier densities of the order of $10^{12} - 10^{13}$ cm$^{-2}$ even without intentional doping. This mechanism of polarization leads to unprecedented high power densities and high current drive capability that are one order of magnitude higher than their silicon or GaAs counterparts [6, 7]. The development of new generations of AlGaN/GaN field-effect transistors (FETs) requires low gate leakage and superior pinch-off characteristics, specifically at elevated temperatures for high temperature microwave power electronics [8]. These properties directly impact the device drain breakdown voltage, radio frequency (RF) performance, and noise figure. In the past, several groups have attempted to achieve gate leakage suppression and superior pinch-off characteristics by using the metal-insulator-semiconductor FETs (MISFETs) [9, 10] or metal-oxide-semiconductor FETs (MOSFETs) [11] device approach. However, the performance level of all these insulated gate devices is well below that of the state-of-the-art AlGaN/GaN HFETs. Recently Khan et al. [12] reported the dc characterization results of AlGaN/GaN metal-insulator-semiconductor heterostructure field-effect transistors (MISHFETs) on sapphire substrates. The built-in channel of MISHFET is formed by the high density 2-DEG at the AlGaN/GaN interface as in regular AlGaN/GaN HFETs. However, in contrast to HFETs, the metallic gate is isolated from AlGaN barrier layer by a thin Si$_3$N$_4$ film. This insulator layer provides extremely low gate leakage current and allows for a large negative to positive gate voltage swing (GVS) [12]. Thus MISHFET combines the advantages of the MIS structure that suppresses the gate leakage current and AlGaN/GaN heterointerface, which provides high-density high-mobility 2-DEG channel. Although piezoelectric polarization results in large values of sheet carrier density, it also gives rise to charged surface states within the device. These surface states are considered responsible for DC to RF current collapse or dispersion, because these electron traps act as a negatively charged virtual gate and limit maximum current available during microwave operation. Good insulator can passivate these surface states and also reduce gate leakage. Thus, the same dielectric can be used both as a gate insulator as well as the surface passivation layer [13]. The MISHFET approach also allows for application of high positive gate voltages to further increase the sheet carrier density in the 2-DEG channel and hence the device peak currents. These features make MISHFETs extremely promising for high power microwave applications. However, there are many milestones to be achieved, and the work in this field is far from complete. Physics-based analytical modeling, which reflects the mechanism of device operation, is an essential requirement to fully explore the performance enhancements of MISHFET.

To characterize and optimize the device performance, an accurate charge control relation between 2-DEG sheet carrier density $n_s$ and the controlling gate voltage $V_g$ is desirable. Various models...

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BROADBAND PARALLEL-COUPLED QUADRATURE COUPLER WITH FLOATING-POTENTIAL GROUND PLANE CONDUCTOR

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ABSTRACT: The design of a parallel-coupled directional coupler, which has a broadband performance, is presented. A floating-potential ground plane conductor is used underneath the coupled region to relax the requirement for a narrow slot between the coupled lines and to improve the isolation and return loss performance of the device. The simulated and measured results show that the designed coupler exhibits a coupling of 3 ± 1 dB across the band 4–10.2 GHz, when the spacing between the coupled lines is 0.13 mm, which is easy to manufacture. This broadband coupling is accompanied by an isolation and return loss which are better than 15 dB across the band 3.1–10.6 GHz. The simulated and measured results show good phase stability and flat group delay across the ultra wideband. The designed device has a compact size and easy to manufacture. The simulated and measured results show an improved performance concerning the return loss and the isolation compared with the previous design [9], where the slotted ground plane was used without adding the floating-potential conductor.

2. DESIGN

The configuration of the proposed directional coupler is shown in Figure 2. The two parallel-coupled microstrip lines are located at the top layer, whereas the ground plane is located at the bottom layer. There is a rectangular slot made at the ground plane underneath the coupled lines. A floating-potential conductor is added to cover most of the slot in the ground plane.

Since a pair of coupled lines over a ground plane is actually a three-conductor transmission line, it can support two different modes of propagation. These modes, which are called the even- and odd-mode, have different characteristic impedances. For the microstrip transmission lines, the dielectric medium is not homogeneous. A part of the field extends into the air above the substrate. This fraction is different for the two modes of the coupled lines. Consequently, the phase velocities, the effective dielectric constants, and the impedances are not equal for the two modes. The performance of the coupled microstrip lines can be approximated using the quasi-static analysis. In this case, the properties of the coupled lines can be determined from the self- and mutual capacitances for the lines. These capacitances depend on the distribution of the electromagnetic field in the structure which, in turn, depends on the dimension of the structure. Distribution of the electromagnetic field in the two modes of operation for the structure under investigation is shown in Figure 2.

With a slot in the ground plane underneath the coupled lines, both the even- and odd-mode capacitances of the coupled lines are decreased substantially. With an additional separated rectangular
conductor inserted in the slot under the coupled lines, the odd-mode capacitance is increased, while the even-mode capacitance is not changed. This estimation can be justified as follows: for the even-mode, there is a virtual magnetic wall between the two coupled lines as shown in Figure 2(a). Therefore, the added conductor continues to be effectively in a floating-potential situation. It has negligible effect on the distribution of the electromagnetic fields of the coupled lines for this mode. Hence, the capacitance in this case is almost equal to its value when there is no conductor to cover the slot in the ground plane. For the odd-mode and due to symmetry of the structure shown in Figure 2(b), the centre of the floating-potential conductor is effectively at zero voltage as the electric wall passes through the middle of this conductor, which means that it is virtually short circuited to the ground. Hence, the effective capacitance at this mode is equal to the capacitance of the conventional parallel-coupled microstrip lines, assuming that the slot $s_2$ in the ground plane is much less than the width of the floating-potential conductor.

Compared with the structure proposed in [9], the presence of a floating-potential conductor, which is inserted in the slot made in the ground plane underneath the coupled lines, offers a compensation for the reduction in the odd-mode capacitance due to the slot in the ground plane. The overall effect of the added floating-potential conductor is that it helps in mode phase-velocity matching, especially in the cases involving low-permittivity substrates for which the slotted ground-plane solution does not provide a satisfactory performance concerning, specifically, the isolation and the directivity.

For each of the two modes of propagation, the capacitance for each of the two lines depends on dimension of the coupled structure. The relation between them can be found using the conformal mapping technique [10, 11] and the image theory [9, 12]. Assuming that the coupled lines are symmetrical, the results are shown in the following equations.

$$C_e = \frac{\varepsilon_r \varepsilon_o w_c q}{h \sqrt{h^2 + 0.25w_e s_1}}$$ (1)

$$C_o = \frac{\varepsilon_r \varepsilon_o w_c q}{h \sqrt{h^2 + 0.25w_o s_1}} + 2\varepsilon_r \varepsilon_o K(k)$$ (2)

where $C_e$ and $C_o$ are the capacitance values per unit length of the even- and odd-mode, respectively, $w_e$: width of each of the coupled lines, $s_1$: the spacing between the coupled lines, $w_o$: width of the slot at the ground plane, $\varepsilon_r$: permittivity of the air, $h$ and $\varepsilon_o$ are thickness and dielectric constant of the substrate, respectively, $K(k)$ is the first kind elliptical integral and $K'(k) = K(\sqrt{1-k^2})$.

Concerning accuracy of (1), caution should be exercised when using it as it was noticed via simulations using Ansoft HFSSv10 that (1) gives reasonable results only when $s_1/h > 0.1$ and $w_e/h > 0.1$. This condition was noticed in the design presented in this letter.

The parameters $q$, $k_1$, and $k_2$ in (1, 2) are functions of the coupler dimension as shown in the following equations, which are the result of the conformal mapping and image techniques [9–12];

$$q = 1 + 1.12(\sqrt{h^2 + 0.25(w_e - s_1)^2/w_e})^0.82$$ (3)

$$k_1 = \frac{s_1}{s_1 + 2w_e}$$ (4)

$$k_2 = \frac{\pi w_o}{4h} \csc \left( \frac{\pi(w_o + s_1)}{4h} \right)$$ (5)

The characteristic impedance of each of the two lines at the even-mode ($Z_{oe}$) and the odd-mode ($Z_{o}$) can be found using the relation [13];

$$Z_{oe} = 1/(\varepsilon_r \varepsilon_o C_o C_e)$$ (6)

$$Z_{o} = 1/(\varepsilon_r \varepsilon_o C_o C_e)$$ (7)

where $C_{oe}$ and $C_{o}$ are the even- and odd-mode capacitances, respectively, when the dielectric is replaced by air and $v_o$ is the velocity of light.
Concerning the coupler’s length $Z$ values of 2306.

To validate the presented design method, a 3-dB directional coupler aimed for the operation within the ultra wideband range.

The measured and simulated insertion loss, return loss, and isolation of the directional coupler are shown in Figure 3. The results indicate an almost flat group delay with very low peak-to-peak variation, which is about 0.03 ns across the band 3–10 GHz.

The simulated group delay for the two output ports are shown in Figure 4. The results indicate an almost flat group delay with very low peak-to-peak variation, which is about 0.03 ns across the band 3–10 GHz.

4. CONCLUSION

The design of a parallel-coupled directional coupler, which has a broadband performance, has been presented. A floating-potential ground plane conductor is used underneath the coupled region to relax the requirement for a narrow slot between the coupled lines and to improve the isolation and return loss performance of the device.

The simulated and measured results have shown that the designed coupler exhibits a coupling of 3 ± 1 dB across the band 4–10.2 GHz, when the spacing between the coupled lines is 0.13 mm, which is easy to manufacture. This broadband coupling is accompanied by an isolation and return loss which are better than 15 dB across the band 3.1–10.6 GHz. The simulated and measured results have shown that the developed device has good phase

Figure 3 The measured and simulated insertion loss, return loss, and isolation of the directional coupler. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

Assuming that the coupler is required to have CF coupling factor, the even- and odd- mode characteristic impedances to achieve this value of the coupling are calculated as follows [13];

$$Z_{ee} = Z_0 \left(\frac{1 + CF}{1 - CF}\right)^{0.5}$$  \hspace{1cm} (8)

$$Z_{oe} = Z_0 \left(\frac{1 - CF}{1 + CF}\right)^{0.5}$$  \hspace{1cm} (9)

where $Z_0$ is the characteristic impedance of the microstrip ports of the coupler.

If $Z_0 = 50 \Omega$ and the coupling factor CF is 0.707 (or 3 dB) then values of $Z_{ee}$ and $Z_{oe}$ can be calculated from (8, 9) and are given as follows: $Z_{ee} = 120.5 \Omega$ and $Z_{oe} = 20.7 \Omega$. Using (1–5), it is possible to find the required dimension for the coupled lines. Concerning the coupler’s length $l$, it is chosen to be quarter of the effective wavelength at the centre frequency of operation.

3. RESULTS

To validate the presented design method, a 3-dB directional coupler aimed for the operation within the ultra wideband range (3.1–10.6 GHz) was designed and tested. Rogers RO4003C (with $\varepsilon_r = 3.38, h = 0.813 \text{ mm}, \text{ loss tangent} = 0.0027$) was used as a substrate for the manufacturing process.

Using the proposed method with the help of the optimization capability of the software HFSSv10, dimensions of the designed directional coupler ($w_c, w_e, l$, and $s_1$) are equal to 1.4, 5.6, 6.2, and 0.13 mm, respectively. The gap between the floating-potential conductor and the rest of the ground plane ($s_2$) was chosen to be 0.4 mm, which is mush less than width of the floating-potential conductor as required by the assumption made in the design procedure (Section 2).

The value of the spacing between the coupled lines ($s_1$) is practical and manufacturable, assuming the use of the general milling and chemical lithography machines. The design of a parallel-coupled directional coupler using the conventional method, i.e. without the floating-potential conductor and slot in the ground plane requires the spacing to be less than 0.001 mm to achieve the 3-dB coupling factor. It is an impractical value and any error due to the manufacturing process degrades the performance to a large extent. The use of the floating-potential ground plane conductor in the design presented in this letter relaxed the spacing requirement between the coupled lines and gave more flexibility in choosing the dimensions.

The return loss, coupling, and isolation of the designed couplers were verified using the software Ansoft HFSSv10 and measured using a vector network analyzer. Figure 3 shows the simulated and measured amplitudes of the scattering parameters for the developed 3-dB coupler. It is clear that the designed coupler features broadband characteristics. The coupling is 3 ± 1 dB across the band 4–10.2 GHz. The isolation and the return loss are better than 15 dB across the band 3–11 GHz. These results indicate a better performance when compared with the previous design [9] which has a slotted ground plane without the floating-potential conductor, where the return loss and isolation are about 13 dB across the band 3–10 GHz.

Concerning the phase performance of the manufactured coupler, Figure 4 reveals that the phase difference between the output ports is 90°± 5° over the band 3–10.6 GHz. This result exposes a quadrature broadband coupler with good phase stability.

In addition to the amplitude and phase balances, the demand for a constant group delay is very crucial for the broadband couplers. The simulated group delay for the two output ports are shown in Figure 4. The results indicate an almost flat group delay with very low peak-to-peak variation, which is about 0.03 ns across the band 3–10 GHz.

Figure 4 Phase and group delay of the designed directional coupler. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
stability and flat group delay across the band 3 GHz. The designed device has a compact size with a dimension of 30 mm × 20 mm.

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CIRCULAR POLARIZATION SQUARE-SLOT ANTENNA FOR DUAL-BAND OPERATION
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ABSTRACT: The design of a printed slot antenna with dual-band circular polarization (CP) characteristic is presented. The proposed antenna is excited by an L-shaped strip with a taper end, connected in series to a microstrip-line-fed located along the diagonal line of the square-slot. Wide impedance and 3 dB axial-ratio bandwidth (CP bandwidth) are measured and the proposed design is suitable for 802.11 a/b/g and WCDMA operations. © 2008 Wiley Periodicals, Inc. Microwave Opt Technol Lett 50: 2307–2309, 2008; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.23633

Key words: dual-band; circular polarization; square-slot antenna; genetic local search algorithm

1. INTRODUCTION
In recent years, considerable concern has arisen over the design of microstrip antennas with CP radiation for wireless communication systems. One of the advantages is the reduction of multipath interference, which is environmental-dependent and difficult to be solved completely. Furthermore, every odd reflection will become an opposing rotation direction of the electric field vector (E). Hence, the CP antenna, being potentially good at rejecting reflection signals, is able to eliminate the arising multipath effects [1]. For a CP radiator, a resonator must simultaneously excite two orthogonal E vectors of equal amplitude and in-phase quadrature. Therefore, a dual-band CP design is more complicated than that required for dual-band linear polarized design. The design of dual-band CP operation via a single probe feed includes; the use of two stacked patches [2], slotted circular patch [3], combination of dielectric resonator antenna (DRA), and microstrip patch [4]. However, relatively few CP antennas embedded with wide slot for broadband and dual-band operation are available in the open literature. Besides the ability to provide greater bandwidth when wide slot is used as the radiating element, the printed slot antenna also offers bidirectional radiation and possesses greater manufacturing tolerances as compared with a normal microstrip patch antenna.

In this letter, a novel microstrip-fed square-slot CP antenna with broadband and dual-band characteristic is proposed. CP radiation is achieved by using an L-shaped strip with a taper end to excite two orthogonal radiation fields and in-phase quadrature. The genetic local search (GLS) algorithm [5] is initially used to determine the dimensions of the proposed antenna with optimum impedance matching and CP bandwidth, followed by employing the commercial software IE3D to predict the performances of the antenna designed by the GLS. The measured results successfully demonstrated a single-feed dual-band CP design operating at 2390 and 5245 MHz, which is suitable for 802.11 a/b/g (2400–2485, 5150–5350 MHz) and WCDMA (2500–2690 MHz) applications. Furthermore, the lower band demonstrated wide impedance and CP bandwidth of around 39 and 40%, respectively.

2. ANTENNA DESIGN
The key parameters of the proposed dual-band CP antenna as depicted in Figure 1 are investigated using commercial simulation software based on the method of moments. A square slot of side r1 = 61 mm is etched on the ground plane of a FR4 substrate with thickness h = 1.6 mm and relative dielectric permittivity εr = 4.4. The L-shaped strip with a taper end is fabricated directly opposite the square slot. Note that l is the length of the strip along the x-axis measured from the centre of the square slot, while w refers to its width. A thin microstrip line with length l and width w, acts as an impedance transformer between the taper end of the L-shaped strip and the 50-Ω microstrip line along the y-axis. l refers to the horizontal length of the taper.

3. EXPERIMENTAL RESULTS
A prototype of the proposed antenna with dimensions; l = 34.7 mm, w = 9.4 mm, l = 4.5 mm is initially studied. Its return loss
Analytical closed-form solutions for different configurations of parallel-coupled microstrip lines

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Abstract: Closed-form solutions are presented for the even- and odd-mode impedances of the parallel-coupled microstrip lines with different configurations. The first set of equations is for the conventional-coupled microstrip lines. The second set is for the parallel-coupled microstrip lines with a slotted ground plane and the third set is for the parallel-coupled microstrip lines with a slotted ground plane and a floating-potential conductor added to cover most of the slot in the ground plane. The solution is achieved using the conformal mapping technique, which also considers the effect of any fringe capacitor that has an impact on the impedance of the coupled lines. To validate the accuracy of the presented equations, a comparison is made with the available empirical solution (only for the conventional-coupled microstrip lines), a numerical solution (using Galerkin’s technique) and the full-wave electromagnetic solution using the software HFSS. The result of the comparison indicates the high accuracy of the presented equations over a wide range of design parameters. The difference between the calculated values using the derived equations and those from the full-wave electromagnetic analysis is <5% for most of the cases. To validate the derived equations experimentally, they are used to design a 3 dB coupler. The result of measurements on the manufactured coupler confirms the accuracy of the presented equations.

1 Introduction

The parallel-coupled microstrip lines have been used extensively as basic elements for directional couplers, filters, phase shifters and a variety of other useful circuits since mid-1950s [1–3]. Numerous papers have been published describing the fabrication of the conventional parallel-coupled lines. Concerning their design, numerical methods [4, 5] or approximate analytical methods [6–9] were proposed. Numerical studies can provide accurate results, but they are not convenient in the actual uses, such as in the design optimisation. On the other hand, analytical studies result in closed-form expressions which are suitable for analysis and design, but the accuracies depend to a large extent on the involved approximations. In another method to obtain accurate closed-form equations, the empirical method was adopted in [10, 11] by curve fitting of full-wave numerical analysis. According to [12, 13], the equations presented in [11] are the most accurate ones available in the literature for the conventional-coupled microstrip lines. Although the empirical equations in [11] give accurate estimation for the impedances of the coupled lines over a wide range of design parameters, they are lengthy and do not give a clear physical meaning of the effect of each parameter. Because the empirical equations in [11] are so lengthy and contain many intermediate parameters, the authors of [11] needed to submit a correction paper [14] a year later to correct some of their equations.

The conventional parallel-coupled microstrip lines are usually used in applications that cover the loose coupling range, that is, >10 dB. In several applications, a tight coupler with a 3 dB coupling is required. Using a parallel-coupled structure, it is difficult to realise this value of coupling as a very narrow spacing is required, which makes
the manufacturing process difficult. In [15], a slotted ground plane was suggested as a solution to increase the level of coupling in the parallel-coupled directional couplers. The same solution was adopted for the design of filters [16–18]. The slotted ground plane is used to tune the even/odd phase velocities. Thus, it provides a tight coupling with a relaxed requirement on the physical dimensions of the coupled lines in comparison with the conventional parallel-coupled microstrip lines. In another development, it was noticed that the presence of a floating-potential conductor, which is inserted in the slot made in the ground plane underneath the coupled lines, offers a compensation for the reduction in the odd-mode capacitor because of the slot in the ground plane [19]. The slotted ground plane with a floating-potential conductor was used to design bandpass filters and phase shifters which need a high level of coupling [20, 21].

In the above mentioned papers [15–21], the design of the coupled lines with slotted ground plane (with or without the floating-potential conductor) was accomplished using the commercial full-wave electromagnetic solvers. The full-wave analysis is very involving and does not give any physical insight into the effect of each design parameter on the overall operation of the coupled structure. The full-wave solver was used to find the parameters against frequency behaviour of the structure. The performance of the structure is fully unpredictable until the optimised solutions are achieved through a trial-and-error iterative process. Hence, the simulation methods without any design guidelines are time consuming and may not lead to the optimum design. This paper overcomes this limitation by deriving the design equations that can be used directly to find the even- and odd-mode impedances depending on the physical dimensions of the structure. The quasi-static analysis with the help of the conformal mapping technique is adopted in this paper. Moreover, simple design equations for the conventional parallel-coupled microstrip lines are also derived and compared with the empirical equations in [10–14].

## 2 Quasi-static analysis of the parallel-coupled microstrip lines

The three different configurations of the parallel-coupled microstrip lines under investigation in this paper are shown in Fig. 1. The top layer contains the two coupled lines, whereas the ground plane is located at the bottom layer. The structure #1 shown in Fig. 1a is for the conventional parallel-coupled microstrip lines. Fig. 1b shows the structure #2; parallel-coupled microstrip lines with a slotted ground plane. There is a rectangular slot made at the ground plane underneath the coupled lines. The structure #3 shown in Fig. 1c is similar to the structure #2, except that a floating potential conductor is added to cover most of the slot in the ground plane.

![Figure 1 Three different configurations of the parallel-coupled microstrip lines](image)

Since a pair of coupled lines over a ground plane is actually a three-conductor transmission line, it can support two different modes of propagation. These modes have different characteristic impedances. For the microstrip transmission lines, the dielectric medium is not homogeneous. A part of the field extends into the air above the substrate. This fraction is different for the two modes of coupled lines. Consequently, the phase velocities, the effective dielectric constants and the impedances are not equal for the two modes. The performance of the coupled microstrip lines can be approximated using the quasi-static analysis. In this case, the properties of the coupled lines can be determined from the self- and mutual capacitances for the lines.

The structure presented in Fig. 1 can be fully analysed using the even- and odd-mode of operation. Distribution of the electric field lines between the coupled lines for these two modes is shown in Fig. 2.

Assuming a quasi-transverse electromagnetic propagation, the electrical characteristics of the coupled lines can be completely determined from the effective per unit length capacitances and the phase velocity of the lines [22]. Therefore the structures shown in Fig. 3 can be used to derive the required equations for the even- and odd-mode impedances.

The capacitance for the even-mode \(C_{ea}\) and for the odd-mode \(C_{oa}\) for the three configurations shown in Figs. 3a–3c can be determined as follows

\[
C_{1e} = C_d + C_{f3}; \quad C_{1o} = C_6 + 2C_a + 2C_d \quad (1)
\]

\[
C_{2e} = C_G + C_{f2}; \quad C_{2o} = C_G + 2C_a + 2C_d \quad (2)
\]
Figure 2 Distribution of the electric field lines for the two fundamental modes of the parallel-coupled lines in the configurations
a #1
b #2
c #3

Figure 3 Self and mutual capacitances for the two fundamental modes of the parallel-coupled lines in the configurations
a #1
b #2
c #3
\[ C_{3e} = C_G + \frac{C_{GS} C_S}{C_{GS} + C_S} + C_f; \]
\[ C_{3o} = C_G + 2C_a + 2C_d + \frac{C_{GS} C_S}{C_{GS} + C_S} \]  

(3)

It is to be noted here that although the same notation was used for the capacitance \( C_S \) between the floating-potential conductor and the ground plane for the even- and odd-mode of the structure shown in Fig. 3, the real effect of it on the odd-mode is negligible. This comes from the fact that for the odd-mode and due to symmetry of the structure, the centre of the floating-potential conductor is effectively at zero voltage, which means that it is virtually short circuited to the ground, and hence the effective value of the last term of \( C_{3o} \) in (3) is equal to \( C_{GS} \).

The characteristic impedance of each of the coupled lines at the even-mode \( (Z_{oe}) \) and the odd-mode \( (Z_{0o}) \) can be found using the relation [14]

\[ Z_e = \frac{1}{v_0 \sqrt{C_{ae} C_e}}; \quad Z_o = \frac{1}{v_0 \sqrt{C_{ao} C_o}} \]

(4)

\[ C_{1e} = 2\varepsilon_0 \left[ \varepsilon_i \frac{K(k_1)}{K'(k_1)} + \frac{K(k_2)}{K'(k_2)} \right]; \]
\[ C_{1o} = 2\varepsilon_0 \left[ \varepsilon_i \frac{K(k_1)}{K'(k_1)} + \frac{K(k_2)}{K'(k_2)} \right] \]  

(5)

\[ C_{2e} = \varepsilon_0 \left[ \varepsilon_i \frac{K(k_5)}{K(k_7)} + \frac{8K(k_4)}{\sqrt{\varepsilon_i K'(k_6)}} \right]; \]
\[ C_{2o} = \varepsilon_0 \left[ \frac{K(k_5)}{K(k_7)} + \varepsilon_i \frac{K'(k_4)}{K(k_6)} + \frac{K'(k_2)}{K(k_3)} \right] \]  

(6)

The values of the capacitances used in (1)–(4) depend on the dimension of the coupled structure. The relation between them can be found using the conformal mapping technique [23] and the Schwartz–Christoffel method. The results of the conformal mappings for the three configurations under investigation after a series of transformations are given in the following equations

\[ C_{1e} = 2\varepsilon_0 \left[ \varepsilon_i \frac{K(k_1)}{K'(k_1)} + \frac{K(k_2)}{K'(k_2)} \right]; \]
\[ C_{1o} = 2\varepsilon_0 \left[ \varepsilon_i \frac{K(k_1)}{K'(k_1)} + \frac{K(k_2)}{K'(k_2)} \right] \]  

(5)

\[ C_{2e} = \varepsilon_0 \left[ \varepsilon_i \frac{K(k_5)}{K(k_7)} + \frac{8K(k_4)}{\sqrt{\varepsilon_i K'(k_6)}} \right]; \]
\[ C_{2o} = \varepsilon_0 \left[ \frac{K(k_5)}{K(k_7)} + \varepsilon_i \frac{K'(k_4)}{K(k_6)} + \frac{K'(k_2)}{K(k_3)} \right] \]  

(6)

Figure 4 Comparison between the calculated odd- and even-mode characteristic impedances for the conventional-coupled microstrip lines using the proposed equations and three other methods

- Odd-mode with \( \varepsilon_r = 3.38 \)
- Even-mode with \( \varepsilon_r = 3.38 \)
- Odd-mode with \( \varepsilon_r = 9.8 \)
- Even-mode with \( \varepsilon_r = 9.8 \)
The design parameters used in (8)–(14) are shown in Fig. 1. Their definitions are as follows.

- \( s \) is the width of each of the coupled lines,
- \( w_c \) is the width of each of the coupled lines,
- \( w_s \) is the width of each of the coupled lines,
- \( w_c \) is the width of each of the coupled lines,
- \( w_s \) is the width of each of the coupled lines,
- \( h_c \) is the spacing between the coupled lines,
- \( h_s \) is the spacing between the coupled lines,
- \( w_c \) is the width of each of the coupled lines.

The design parameters used in (8)–(14) are shown in Fig. 1. Their definitions are as follows.

\[ k_1 = \tanh\left( \frac{\pi w_c}{4h} \right) \tanh\left( \frac{\pi (w_c + s_1)}{4h} \right); \]
\[ k_2 = \tanh\left( \frac{\pi w_c}{4(h + \pi w_c)} \right) \tanh\left( \frac{\pi (w_c + s_1)}{4(h + \pi w_c)} \right) \]
\[ k_3 = \tanh\left( \frac{\pi w_c}{4h} \right) \coth\left( \frac{\pi (w_c + s_1)}{4h} \right); \quad k_4 = \frac{w_c}{w_c + s_1} \]
\[ k_5 = \frac{1 + \exp[-\pi(w_c - s_1)/(2h)]}{1 + \exp[-\pi(w_c - s_1 - 2w_c)/(2h)]}; \]
\[ k_6 = \tanh\left( \frac{\pi w_c}{4(h + d)} \right) \tanh\left( \frac{\pi (w_c + s_1)}{4(h + d)} \right) \]

Figure 5 Comparison between the calculated odd-mode characteristic impedances for the configuration #2 using the proposed equations and two other methods

- \( w_c/h = 4, \varepsilon_r = 3.38 \)
- \( w_c/h = 8, \varepsilon_r = 3.38 \)
- \( w_c/h = 4, \varepsilon_r = 9.8 \)
- \( w_c/h = 8, \varepsilon_r = 9.8 \)

\[ d = \begin{cases} \frac{w_c}{(w_c - s_1)/2} & \text{if } w_c > (w_c - s_1)/2 \\ \frac{w_c}{(w_c - s_1)/2} & \text{if } w_c < (w_c - s_1)/2 \end{cases} \]
\[ k_7 = \frac{s_1}{s_1 + 2w_c}; \quad k_8 = \frac{\tanh(\pi s_1/(4h))}{\tanh(\pi (s_1 + 2w_c)/(4h))} \]
\[ k_9 = \frac{[1 + \exp(-\pi s_1/(2h))]}{[1 + \exp(\pi (w_c - 2w_c - s_1 - 2d))/(2h)]}; \]
\[ k_{10} = \frac{w_c}{w_c - 2d}; \quad k_{11} = \frac{\sinh(\pi s_1/(4h))}{\sinh(\pi (s_1 + 2w_c)/(4h))} \]
between the added floating-potential conductor and the ground plane, \( e_r \) the permittivity of the air, \( h \) and \( r \) the thickness and dielectric constant of the substrate, respectively, \( K(k) \) the first kind elliptical integral and \( K'(k) = K(\sqrt{1 - k^2}) \).

It is to be noted that it might be preferable to present one set of equations that can be used for the three configurations. However, one general set of equations would be complicated and very lengthy. It was noticed by the author that dealing with each configuration as a separate case helped to use some reasonable and practical assumptions which made the analysis easy and the final solution simple and accurate. The assumptions used to derive (5)–(14) are explained hereafter. Concerning \( C_{f1} \), the effective distance between the microstrip line and ground plane was assumed to be equal to the average path of the fringe field \((\pi w_s + h)\). The conformal mapping was then used to calculate its value following the same steps used for \( C_{g1} \) except that \((\pi w_s + h)\) is used instead of \( h \). The method adopted here to calculate the effect of the fringe capacitor is different from the empirical method used in [12]. It is worthwhile to mention here that the conformal mapping used to find \( C_{g1} \) does not include the part of field which extends into the air above the dielectric. Hence, the inclusion of \( C_{f1} \) in the design equations is necessary to obtain accurate results. For the fringe capacitor \( C_{f2} \), the average path of the fringe field \((d + h)\), where \( d \) is given in (11), was used in the conformal mapping.

In (6) and (7), it was assumed that width of the slot in the ground plane \((w_s)\) is equal to, or larger than, the width of the substrate \((b)\). This assumption is reasonable because if the slot in the ground plane is required to have a considerable impact on the performance then it should be larger than the width of the ground plane, otherwise the effect will be negligible and the equations of the conventional parallel-coupled microstrip lines (5) can be used. For this reason, the effect of the slotline mode across the slotted ground plane is negligible, and hence it is not included in the calculations for configuration #2. In deriving (7), the gap in the ground plane of the configuration #3 \((s_2)\) was assumed to be equal to, or less than, the width of the substrate \((b)\). This means a significant effect for the slotline mode as shown in Fig. 2c, and hence \( C_{g3} \) shown in Fig. 3c is included in the calculations for configuration #3.

Figure 6 Comparison between the calculated even-mode characteristic impedances for the configuration #2 using the proposed equations and two other methods

\begin{itemize}
  \item[a] \( w_s/h = 4, \quad e_r = 3.38 \)
  \item[b] \( w_s/h = 8, \quad e_r = 3.38 \)
  \item[c] \( w_s/h = 4, \quad e_r = 9.8 \)
  \item[d] \( w_s/h = 8, \quad e_r = 9.8 \)
\end{itemize}
To validate the derived equations, they were used to calculate the even- and odd-mode impedances for a wide range of design parameters and the results were compared with three different methods: the available empirical equations (for the conventional-coupled microstrip lines) [11–14], a numerical calculation by solving a set of integral equations for the bound charges using Galerkin’s technique, with a piecewise constant approximation for the charge distribution [24] and the full-wave electromagnetic analysis using the software Ansoft HFSSv10, which is based on the finite-element method.

Concerning the conventional-coupled microstrip lines, the result for a practical range of values for \( w_c/h = 0.5, s_1/h = 0.5, \epsilon_r = 3.38 \) is shown in Fig. 4 assuming two values for the dielectric constant (3.38 and 9.8). The results shown in Fig. 4 reveal the accuracy of the presented equations (5) in spite of their simplicity compared with the empirical equations given in [11–14]. It is to be mentioned here that those empirical equations suffer from a high error (>20%) when they are used for a narrow spacing \((s_1/b < 0.1)\) and a narrow strip width \((w_c/b < 0.1)\). The equations presented in this paper were checked for the case of extremely narrow spacing \((s_1/b \approx 0.0001)\) and they gave accurate results when compared with the full-wave electromagnetic solution (<5% difference).

For the case of the coupled lines with slotted ground plane (configuration #2), there are no equations available in the literature to compare with the presented formulae; therefore the comparison was made with the results of the numerical solution using Galerkin’s technique and the full-wave electromagnetic solution using HFSS. The results shown in Figs. 5 and 6 indicate the accuracy of the method for a wide range of design parameters. The effect of the slotted ground plane can be explained by comparing the results shown in Figs. 5 and 6 with those in Fig. 4. Making a slot in the ground plane decreases the values of the even- and odd-mode capacitances, and hence increases the values of the even- and odd-mode impedances. The effect of the slot on the even-mode impedance is much greater than its effect on the odd-mode impedance. Increasing the size of the slot in the ground plane has a limited effect on the odd-mode impedance, especially when \( w_c/b < 1 \); compare Figs. 5a and 5b or Figs. 5c and 5d, whereas it has a significant effect on the even-mode impedance; compare Figs. 6a and 6b or Figs. 6c and 6d. Adjusting the size of the slot enables the designer to achieve the required values.

Figure 7 Comparison between the calculated odd-mode characteristic impedances for the configuration #3 using the proposed equations and two other methods

- For the case of the coupled lines with slotted ground plane (configuration #2), there are no equations available in the literature to compare with the presented formulae; therefore the comparison was made with the results of the numerical solution using Galerkin’s technique and the full-wave electromagnetic solution using HFSS. The results shown in Figs. 5 and 6 indicate the accuracy of the method for a wide range of design parameters. The effect of the slotted ground plane can be explained by comparing the results shown in Figs. 5 and 6 with those in Fig. 4. Making a slot in the ground plane decreases the values of the even- and odd-mode capacitances, and hence increases the values of the even- and odd-mode impedances. The effect of the slot on the even-mode impedance is much greater than its effect on the odd-mode impedance. Increasing the size of the slot in the ground plane has a limited effect on the odd-mode impedance, especially when \( w_c/b < 1 \); compare Figs. 5a and 5b or Figs. 5c and 5d, whereas it has a significant effect on the even-mode impedance; compare Figs. 6a and 6b or Figs. 6c and 6d. Adjusting the size of the slot enables the designer to achieve the required values.
for the even- and odd-mode impedances, and hence accomplish the required value of the coupling factor.

The last structure under investigation is configuration #3, which is shown in Fig. 1c. The results of the calculation for the even- and odd-mode impedances using the derived equations are presented in Figs. 7 and 8 for a wide range of design parameters. The values of the impedances as a result of using Galerkin’s technique and the full-wave electromagnetic solution are also shown in Figs. 7 and 8. These figures depict the accuracy of the presented equations.

As can be seen from Figs. 7 and 8, the addition of a conductor to cover most of the slot in the ground plane decreases both the even- and odd-mode impedances. However, the effect of the floating-potential conductor on the odd-mode is greater than its effect on the even-mode capacitor, especially when $w_s/h \gg 1$. The addition of the conductor restores the value of the odd-mode impedance obtained for the conventional microstrip lines. The values of the odd-mode impedance for the configuration #3 in Fig. 7 are almost equal to their counterpart for the conventional-coupled microstrip lines in Fig. 4. For the even-mode, the added conductor decreases the impedance only by a small value when compared with configuration #2. Hence, the even-mode impedance can still be controlled by the width of the slot in the ground plane ($w_s$).

### 3 Design of a 3 dB coupler

As another step for the validation of the derived equations, a 3 dB directional coupler is designed assuming the three configurations and a comparison is made with the values of the design parameters when using the other three methods. The designed 3 dB coupler using the derived equations in this paper was also manufactured and tested.

Assume that it is required to design a coupler which has $C$ coupling factor. The even ($Z_{oe}$) and odd ($Z_{oo}$) mode characteristic impedances to achieve this value of the coupling are calculated as follows

$$Z_{oe} = Z_o \left(\frac{1+C}{1-C}\right)^{0.5} \quad Z_{oo} = Z_o \left(1+C\right)^{0.5}$$

where $Z_o$ is the characteristic impedance of the microstrip ports of the coupler.
If $Z_0 = 50\,\Omega$ and the coupling factor $C$ is 0.707 (or 3 dB) then the values of $Z_{oe}$ and $Z_{oo}$ can be calculated from (15) and are given as follows: $Z_{oe} = 120.5\,\Omega$ and $Z_{oo} = 20.7\,\Omega$. Using (4)–(14), it is possible to find the required dimension for the coupled lines of the three configurations under investigation. The results of calculation using the derived equations, besides the results from the empirical equation for the conventional-coupled microstrip lines [11–14], Galerkin’s technique and the software HFSS are shown in Table 1. The high accuracy of the presented equations is clear from the table, where the difference between the calculated values when using the derived equations compared with the full-wave electromagnetic analysis, using the software HFSS, is <5% for most of the cases listed in Table 1. Concerning the conventional parallel-coupled microstrip lines, the results listed in Table 1 reveal that the derived equations in this paper are more accurate compared with the empirical equations [11–14] assuming the full-wave analysis of HFSS as a reference for the comparison.

Table 1 also shows the importance of making a slot in the ground plane underneath the coupled lines and then adding a floating-potential conductor to cover most of the slot. The width of the coupled lines and the spacing between them have been relaxed from low and impractical values in the conventional configuration to reasonable and practical values, which can be easily manufactured, in the configuration #3.

A 3 dB directional coupler utilising the configuration #3 and designed using the derived equations was manufactured and tested. Rogers TMM10i (with $e_r = 9.8$, $h = 0.635\,\text{mm}$, loss tangent = 0.0015) was used as a substrate. The values of the design parameters are listed in the last row of Table 1. Concerning the coupler’s length $l$, it is chosen to be quarter of the effective wavelength at the centre frequency of operation. Assuming that it is required to develop a C-band (4–8 GHz) coupler, then the centre frequency is 6 GHz. The overall dimension of the developed coupler including the input/output ports is 20 mm × 30 mm.

The performance of the developed coupler was simulated using the full-wave electromagnetic simulator (HFSSv10) and measured using a vector network analyzer. The

Table 1. Calculated values of the design parameters for the 3 dB coupler using the derived equations compared with other methods

<table>
<thead>
<tr>
<th>Type of the coupled lines</th>
<th>Assumed parameters</th>
<th>This paper</th>
<th>Kirschning and Jansen [11, 14]</th>
<th>Galerkin’s technique</th>
<th>HFSSv10</th>
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<td></td>
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<td>—</td>
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<td>—</td>
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Figure 9 Measured performance of the manufactured 3 dB coupler designed using the derived equations.
simulated and measured return loss, coupling and isolation of the manufactured coupler are in good agreement as shown in Fig. 9. It is clear that the designed coupler features a 3 dB coupling at the design frequency (6 GHz). This value of coupling is in complete agreement with the design value even without the need to do any further adjustment or optimisation. It is worth to mention that Fig. 9 shows a coupling error of $3 \pm 1$ dB, whereas the isolation and return loss are better than 20 dB across the whole C-band.

4 Conclusion

Quasi-static solutions have been presented for the even- and odd-mode impedances of the parallel-coupled microstrip lines with different configurations. The first set of equations is for the conventional-coupled microstrip lines, whereas the second and third sets are for the slotted-ground coupled lines with or without a floating-potential conductor at the slot in the ground plane. The solution has been achieved using the conformal mapping technique. To validate the accuracy of the presented equations, a comparison has been made with the available empirical solution (only for the conventional-coupled microstrip lines), a numerical solution (using Galerkin’s technique) and the full-wave electromagnetic solution using the software HFSS. The result of the comparison has indicated the high accuracy of the presented equations over a wide range of design parameters. To validate the derived equations experimentally, they are used to design a 3 dB coupler. The result of measurements on the manufactured coupler has confirmed the precision of the presented equations.

5 References


A theoretical model is used to investigate the effect of the tapering shape on the performance of the broadside-coupled directional coupler. It is shown that the tapering shape has a significant effect on the overall performance. Concave shapes give a broader bandwidth compared with the convex shapes. The model also shows that a proper choice of the tapering shape can result in a significant improvement in the performance across a wide bandwidth. The theoretical findings of the model are supported by simulations and measurements.

**Key words:** directional coupler; tapering; ultra wideband

1. **INTRODUCTION**

Directional couplers are very important passive microwave devices, which are used to divide signals with appropriate phase. They are widely used in balanced mixers, modulators, and beam forming networks.

To build broadband directional couplers, coupled transmission lines are usually used, such as in Lange [1] and tandem couplers [2]. However, those configurations have some serious limitations, such as the need for wire crossovers and narrow strips, which create manufacturing problems. To avoid these problems, the broadside coupling approach, which requires the use of a double-sided substrate, was utilized [3]. The structure is formed by two microstrip patches separated by a rectangular slot in the common ground plane.

In the last few years, the ultra wideband (UWB) technology has emerged as a hot research area which attracts much attention. As a response to this new emerging technology and to avoid the drawbacks of the previous methods in designing directional couplers, a broadside-coupled structure, which utilizes elliptical shapes for the coupled patches and slots, was proposed [4]. The results presented in [4] indicate a UWB performance. However, details of the design procedure do not show why the elliptical shape was specifically chosen and whether or not other shapes can achieve similar, or even better, performance.

In this article, a theoretical investigation is introduced to show the effect of different tapering shapes on performance of the broadside-coupled directional coupler. To accurately model the utilized structure, the partial reflection theory presented in details in [5–8] and the conformal mapping approach [9] are used.

2. **MODEL**

Configuration of the utilized structure of the multilayer broadside-coupled directional coupler is shown in Figure 1. The device consists of three conductive layers interleaved by two dielectrics. The top conductive layer includes the input port (Port 1) and the direct output port (Port 2). The bottom conductive layer is similar to the top layer but the ports are the coupled output port (Port 3) and the isolated port (Port 4), which has no output power, and thus it is connected to a matched load. The two layers are coupled via a slot made in the mid layer, which forms the ground plane.

To analyze the structure shown in Figure 1, the theory presented in [5–8] is utilized. This theory suggests that the incident signal is partially reflected at every strip of any tapered transmission line because of the mismatch between the consecutive strips. The total reflection coefficient at the input port can then be found by summing differential contributions of the reflection coefficient from each strip in proper phase. To get accurate results, that theory suggests that the strip size should be very small compared with the wavelength.

Referring to Figure 1(b) and assuming that the even-mode impedance at position \( x \) is \( Z_e \) and at position \( (x + dx) \) is \( (Z_e + dZ_e) \), the differential reflection coefficient due to a strip of width \( dx \) is given by [5–8];
\[ dS_{11e} = e^{-2\beta x} \frac{Z_x + dZ_x - Z_x}{Z_x + dZ_x + Z_x} dZ_x = e^{-2\beta x} \frac{dZ_x}{2Z_x} \]

\[ = 0.5e^{-2\beta x} \frac{dZ_x}{dx} (\ln Z_x) dx \]  

(1)

where \( \beta \) is the even-mode phase constant, and \( l \) is length of the tapered structure. Assuming that the substrate supporting the structure has a dielectric constant equal to \( \varepsilon_r \) and the wavelength of operation is \( \lambda \) then the approximate value for \( \beta \) for the structure under consideration is equal to:

\[ \beta = \frac{2\pi \sqrt{\varepsilon_r}}{\lambda} \]  

(2)

Using the integration to add up all the partial reflections across the coupled structure, the total even-mode reflection at the input port (\( S_{11e} \)) is equal to:

\[ S_{11e} = 0.5e^{-\beta l} \int_0^l e^{-2\beta v} (\ln Z_x) dv \]  

(3)

According to the theory of backward quadrature couplers [8], if the input/output ports are perfectly matched, the coupled signal to Port 3 (\( S_{31} \)) is equal to \( S_{11e} \). Equation (3) can be arranged in the following form after introducing the intermediate parameter \( \nu = x - \nu_2 \):

\[ e^{\jmath \beta \nu} S_{31} = e^{\jmath \beta \nu} S_{11} = \int_{\nu_2}^{\nu_1} e^{-2\beta \nu} (\ln Z_x) d\nu \]  

(4)

Knowing that the integration is zero outside the range \( \nu \in [\nu_2, \nu_1] \). This equation represents the Fourier transform of the function \((\ln Z_x)/(2\nu)\) between the domains \( \nu \) and \( 2\beta \). Taking the inverse Fourier transform of (4) results in:

\[ \left( \frac{d(\ln Z_x)}{2\nu} \right) = \int_{-\infty}^{\infty} e^{2\beta \nu} (e^{\jmath \beta \nu} S_{11}) d(2\beta) \]  

(5)

If it is required to achieve \( C \) coupling between the input port and the coupled port (Port 3 in Fig. 1(a)), then \( S_{11} \) across the desired band can be written as [8],

\[ S_{11} = C e^{\jmath \varphi \nu^2 - \beta \nu} \]  

(6)

where a quadrature coupler is assumed here. Substituting from Eq. (6) in Eq. (5) and rearranging:

\[ Z_x(x) = Z_e^{\varphi \nu^2} \]  

(7)

\[ \delta = \frac{2C}{\pi} \int_{-\nu_2}^{\nu_1} e^{\jmath \varphi \nu^2} \sin (2\beta \nu) \nu d\nu \]  

(8)

For the broadside-coupled structure shown in Figure 1, the even-mode impedance at any \( x \) within the range \( x \in [\nu_2, \nu_1] \) can be calculated using the conformal mapping [4, 9] after utilizing the quasi-static approximation:

\[ f_1(x) = \left( \frac{w_x - w_i}{2(C_{024} - 1)} \right)(e^{\nu^2} - 1) + w_i/2 \]  

(11)

\[ f_2(x) = \left( \frac{w_x - w_i}{2(C_{024} - 1)} \right)(e^{\nu^2} - 1) + w_i/2 \]  

(12)

The parameter \( n \) is used to change shape of the tapering as shown in Figure 2. Positive values for \( n \) result in a convex shape, whereas negative values refer to a concave structure. When \( n = 0 \), the structure has a triangular shape, while when \( n = -\infty \), it has a rectangular shape.

Equations (11) and (12) are used to model the structure for \( 0 \leq x \leq \nu_2 \), while, for \( \nu_2 \leq x \leq \nu_1 \), the structure is its image assuming

Figure 2 The tapering shape at different values of the parameter \( n \). [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
In Figure 3, the calculated coupled output for different values of $n$. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

In Figure 4, performance of the optimized exponentially tapered structure is also compared with that of the elliptical shape adopted in [4]. It is clear that the exponential structure achieves a better performance concerning the coupled and direct output at the lower end of the frequency band, while the elliptical shape has a very close to the elliptical shape. This explains the broadband performance of the elliptical structure used in [4]. Referring to Figure 3, it is also obvious that the tapered concave shapes with $-2 \leq n \leq 0$ gives a better performance compared with the rectangular shape used in some published papers, such as in [3], while the convex shape, i.e., $n$ is positive, has the worst performance.

The software CST Microwave Studio was then used to make sure of accuracy of the theoretical modeling proposed in this paper assuming the same substrate, frequency range, and coupling. To increase the degree of freedom in the design, the structure was also analyzed when a different value of $n$ and a different length $l$ are used for the coupled patches and the slot. It was found that the optimum structure has the same value of $n$ for both the coupled patches and the slot and it is $-0.98$, which is very close to the $-0.9$ value obtained via the theoretical model. This proves the reasonable accuracy of the presented approach. Concerning the other design parameters, the optimized values are: $l$ (for the coupled patches) = 4.86 mm, $l$ (for the slot) = 5.3 mm, $w_c = 3.3$ mm, $w_e = 8.6$ mm, $w_l = 0.67$ mm. A prototype with the optimized parameters was manufactured and tested to validate the outcome of the analysis and simulations. The result of measurements concerning amplitude of the $S$ parameters is shown in Figure 4 alongside the simulated result, while the phase performance is shown in Figure 5. These results show a 3 dB quadrature and UWB performance, indicate a good agreement between the simulated and measured values. The simulated and measured coupled power in Figure 4 confirms the theoretical model when compared with the general variation of the coupled power for $n = -0.9$ in Figure 3.

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better performance at the rest of the UWB. Concerning the return loss and isolation, the elliptical shape has a better performance across most of the UWB.

4. CONCLUSION
In this article, a theoretical model has been used to study effect of the tapering shape on performance of the broadside-coupled directional coupler across the ultra wideband range. Simulations and measurements have been used to verify the findings of the model.

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EFFECT OF THE CONDUCTIVE COATING THICKNESS ON PERFORMANCE OF THE MICROSTRIP REFLECTARRAY

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ABSTRACT: Effect of thickness of the conductive coating on performance of the microstrip reflectarray is investigated. The unit cell used in this article is in the form of a cross shaped ring. Dimensions of the utilized cell are chosen such that it resonates at the X-band. The simulated and measured results show that increasing thickness of the conductive layer decreases the losses significantly, reduces the phase slope and shifts the resonant frequency to a higher value, while it has a negligible effect on the phase range. © 2009 Wiley Periodicals, Inc.

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Key words: reflectarray; microstrip antenna; conductor loss

1. INTRODUCTION
The microstrip reflectarray is a high gain antenna which consists of a combination of a flat reflector and a planar phased microstrip array. It uses a suitable phasing scheme to convert a spherical wave produced by its feed into a plane wave, and thus it can be used as an alternative to parabolic reflectors [1, 2]. Despite the considerable development in the design of different configurations of the microstrip reflectarray, it still suffers from a significant deterioration in its efficiency due to the combined effect of dielectric loss, conductor loss and surface wave excitation [3, 4]. It has been shown that the dielectric loss and the conductor loss are much more significant than the losses due to the surface wave excitation [3]. In a recent work [5], it has been shown by simulations and measurements that the dielectric loss is inversely proportional with thickness of the substrate.

In this article, effect of the conductive coating thickness on the total losses, and hence on performance of the microstrip reflectarray is investigated. A single-layer cross shaped ring presented in [6] is used as a reflectarray element for the undertaken investigation. The cross shaped ring has a broad bandwidth compared with other shapes, such as the printed dipoles or patches, and it is easy to manufacture compared with the stacked elements [6]. In the presented investigations, it is shown that thickness of the conductive coating is an important parameter to be considered when trying to maximize the efficiency of the microstrip reflectarray.

2. RESULTS AND DISCUSSION
In the undertaken investigations, the cross shaped ring shown in Figure 1 is utilized [6]. It was designed to operate in the X-band with the centre frequency of 10 GHz. The waveguide model was used to calculate effect of the conductive layer thickness on performance of the unit cell element. Dimensions of the cross shaped ring used in the investigation are: L1 = 7.5 mm, L2 = 1.3 mm, W = 0.2 mm. To compare the simulated and measured results, size of the unit cell was chosen to be equal to dimensions of the standard X-band rectangular waveguide used in the measurements, which is 22 ×11 mm². The substrate used to support the cross element is Rogers RT5880 with εr = 2.2, dielectric loss tangent = 0.0009, and thickness = 3.175 mm. Several values for
3. EXPERIMENTAL RESULTS

The filter is machined using LPKF printed circuit board prototyping machine. Figure 5 shows the photograph of the experimental broadband filter. Figure 6 compares the experimental results of designed filter against the full wave simulation results. Comparison shows a good agreement between them confirming the expected broadband and suppressed second harmonic features. Frequency band of the filter is 3.1 to 6.75 GHz. Maximum insertion loss of the filter is 0.4 dB, and return loss is better than 13 dB. Stop band rejection is better than 25 dB over 7.5 GHz to 12 GHz while the second harmonic (9.9 GHz) of the filter has been suppressed to a level of 30 dB. Filter is compact and size is $12 \times 15 \times 0.78 \text{mm}^3$.

4. CONCLUSION

Using a quarter wave coupled microstrip resonator in a defected ground configuration, a broadband filter from 3.1–6.8 GHz was designed and analyzed. Filter used a short circuited quarter wave resonator for second harmonic suppression. The filter exhibited 0.4 dB insertion loss and 13 dB return loss over the pass band. The results of analysis were confirmed through experiment. Overall dimensions of the filter are $12 \times 15 \times 0.78 \text{mm}^3$.

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MULTI-OCTAVE MICROSTRIP-TO-COPLANAR WAVEGUIDE VERTICAL TRANSITION

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ABSTRACT: A vialess vertical microstrip-to-coplanar waveguide (CPW) transition that covers a multioctave bandwidth is proposed. The proposed transition utilizes the magnetic coupling in a pair of microstrip-to-slotline transitions derived from the microstrip/CPW structure. The presented device is designed following simple design guidelines. The simulated and measured results show that the proposed transition can achieve a six-octave bandwidth. © 2010 Wiley Periodicals, Inc. Microwave Opt Technol Lett 53:187–189, 2011; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.25675

Key words: transition; coplanar waveguide; microstrip

1. INTRODUCTION

As microwave circuits become more compact, new techniques for integration are being utilized. In the modern multilayer technology, the third dimension is utilized for vertical integration to reduce total space and cost. Vertical transitions are thus crucial in the design of multilayer circuits.

Microwave circuits are usually based on planar technologies, such as microstrips and/or coplanar waveguides (CPW), as they provide a compact, lightweight, and low-loss transmission medium. Concerning the multilayer integrated circuits, they require a flexibility to use both microstrip and CPW circuit technologies [1]. Therefore, vertical transitions between microstrip and CPW lines located at different layers are a must to accomplish the much needed flexibility in the design of multilayer circuits. In addition to that, vertical transitions can be used to develop new devices and/or to improve the performance of some of the existing devices [2, 3].

Microstrip-to-CPW vertical transitions are usually designed using either aperture-coupled structures or via-holes. Concerning the via-hole transitions, it was revealed that as the operating frequency increases, the performance of the via-holes is degraded [4]. In addition, their fabrication process is usually difficult and costly as sophisticated tools are needed to minimize their additional losses [5, 6]. To overcome the shortcomings of the via-holes, aperture-coupled vertical transitions can be used [7, 8]. However, the relatively high insertion loss due to the use of compact aperture-coupled transitions in broadband applications, such as the ultra-wideband technology (3.1–10.6 GHz), is still a problem which needs to be solved.

In this article, a microstrip-to-CPW vertical transition is designed by utilizing the magnetic coupling between two 100-$\Omega$ slotlines derived from the 50-$\Omega$ CPW at the bottom layer and two 100-$\Omega$ microstrip lines formed by splitting the 50-$\Omega$ microstrip at the top layer. The design is accomplished following simple design guidelines. The success of the proposed transition is demonstrated via simulations and measurements.

2. DESIGN

The configuration of the proposed vertical transition is shown in Figure 1. In this configuration, the microstrip feeder is assumed to be at the top layer, while the CPW is located at the bottom layer.

The 50-$\Omega$ microstrip line is divided into two similar sections, each having an impedance of 100-$\Omega$, see Figure 1(a). Similarly, the central strip of the CPW at the bottom layer is increased in width so that it forms two slotlines extending in different directions as depicted in Figure 1(b). Thus, the microstrip-to-CPW transition is transformed into a pair of microstrip-to-slotline transitions. In each pair, the microstrip and slotline extend normally beyond each other by a distance of $\lambda/4$, where $\lambda$ is the effective wavelength calculated at the center of the frequency band. This configuration can be considered a magnetic-coupled structure. A signal flowing into each of the microstrip lines projects a strong magnetic field through one of the slotline opening at the other side of the substrate, and thus, it enables the normal slotline propagating mode. Therefore, the signal is efficiently
launched down the two slotlines and eventually via the CPW output.

The strong coupling between each pair of the microstrip and slotlines depicted in Figure 1(c) can be modeled by a transformer with 1:1 turns ratio connecting the two lines. With this modeling, it becomes clear that to have full matching between the 100-Ω microstrip at one side of the transformer and the slotline at the other side, the slotline should also have 100-Ω impedance. The slotline dimension to achieve this value can be calculated using, for example, the closed form solution [9].

To efficiently couple a signal from the microstrip to the slotline across a wide bandwidth, the end of the slotline needs to be compensated with an inductive element, whereas the microstrip needs to be compensated with a capacitive element [2]. The inductive element is chosen in the form of a circular slot, whereas the capacitive element is in the form of a circular disc [2]. The exact location of those elements is shown in Figure 1(c).

It is to be noted that the presented design procedure results in a transition having one transmission pole. To increase the bandwidth, each pair of the circular slots (at the bottom layer) and circular discs (at the top layer) is overlapped in the manner shown in Figure 1(c). This topography adds another transmission pole due to the additional broadside-coupled microstrip–slotline. A parametric analysis can be utilized to find the optimum value of overlapping to achieve the required bandwidth.

3. RESULTS AND DISCUSSION

The validity of the presented design method was tested by building a transition aimed to operate across the band (2–12 GHz). Rogers RT6010 (with □ ≈ 10.2, h = 0.635 mm, and loss tangent = 0.0023) was used for the development. Using the proposed design procedure and with the help of the parametric and optimization capabilities of the software CST Microwave Studio, the dimensions of the transition were found to be: \( r_m = 1.3 \) mm, \( r_s = 2 \) mm, \( s = 0.3 \) mm, and \( w_m = 0.6 \) mm. For the CPW, width of the central conductor is 0.6 mm. The manufactured transition has an overall dimension of 10 × 8 mm\(^2\) excluding the input/output ports. The device was tested via simulation and measurement. Subminiature A connectors were used to connect the manufactured device to the measuring tool.

Figure 2 shows the simulated and measured performance of the designed transition. It is clear from Figure 2 that the simulated and measured results are in good agreement. The presented results reveal that the proposed transition covers the band 1.5–12 GHz assuming the 3-dB insertion loss as a reference. Concerning the return loss at the input/output ports of the transitions, the results in Figure 2 reveal that the return loss at the two ports is more than 20 dB according to the simulations and more than 18 dB according to the measurements across most of the band under investigation. Regarding the insertion loss, it is around 0.1 dB in the simulations and less than 0.4 dB according to the measured results across the band 2.5–11 GHz.
The transition is usually required to have a flat group delay, especially when used in impulse radio systems to minimize the overall system’s distortion. Concerning the proposed transition, Figure 2 shows that it has an almost flat group delay with less than 0.15 ns peak-to-peak variation in the group delay across the investigated band.

Finally, the results shown in Figure 2 resembles bandpass filter (BPF). Thus, the proposed transition can be easily utilized to form, for example, ultra-wideband BPF (3.1–12 GHz) by adjusting values of the design parameters.

4. CONCLUSIONS

Vertical microstrip-to-CPW transition has been presented. The required performance is achieved by transforming the microstrip/CPW structure into a pair of microstrip-to-slotline transitions, which are magnetically coupled. Simple design guidelines are used to design the transition. The simulated and measured results of the proposed transition have shown a bandwidth that extends from 1.5 to 12 GHz.

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APPLICATION OF HYBRID FINITE-DIFFERENCE MODE-MATCHING METHOD TO ANALYSIS OF STRUCTURES LOADED WITH AXIALLY SYMMETRICAL POSTS

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ABSTRACT: In this study, a new hybrid method is proposed to the analysis of waveguide structures loaded with arbitrary configuration of axially symmetrical posts. The method is based on a combination of finite difference frequency domain method, mode-matching technique, and an analytical iterative scattering procedure. The results concerning the accuracy of the proposed technique are discussed. The validity of the proposed approach is verified with the analytical mode-matching method, own measurements, commercial software, and the results published in literature. © 2010 Wiley Periodicals, Inc. Microwave Opt Technol Lett 53:189–194, 2011; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.25644

Key words: hybrid method; FDFD; mode-matching; axially symmetrical

1. INTRODUCTION

The rapid expansion of wireless communication systems has significantly increased the demands for fast and accurate full-wave analysis techniques of microwave components of these systems. The commonly utilized group of such components are waveguide structures, i.e., combline filters [1], resonators [2], multiplexers, and power dividers [3], where cylindrical dielectric or metallic posts are used to obtain desired scattering parameters. The axial symmetry of these objects allows one to limit the analysis to two-dimensional domain corresponding to an angular cross-section of the full three-dimensional object. One of the most efficient and precise analysis techniques of aforementioned structures are the analytical methods, e.g., mode-matching (MM) method [4, 5] or integral equation method [1, 6]. The disadvantage of these approaches is the lack of versatility because they are only applicable to a few regularly shaped waveguide discontinuities. Moreover, the change of the shape beyond the established geometry causes the necessity of a new formulation of the problem. When the axially symmetrical objects with more general shape are taken into account, the more powerful are 2.5-dimensional discrete techniques, e.g., finite difference time domain method (FDTD) [7] or finite element method [8]. However, in the analysis of multiple post complex structures, the analysis using discrete techniques can be inefficient because of the high mesh density involved for accurate results. In this case, the more suitable techniques are hybrid methods [9, 10]. In aforementioned methods, the discrete techniques such as finite element method or FDTD method are used only in limited region surrounding the analyzed structure. In the outer region, the fields are combined with analytical solution of the problem using, e.g., integral equation technique. As a result, the hybrid techniques allows one to achieve higher flexibility, increase the accuracy, and reduce the numerical complexity of the analysis.

In this study, we propose a new hybrid method, which combines the finite difference frequency domain (FDFD) method with the analytical MM technique. The presented approach allows one to analyze scattering from arbitrary set of cylindrical objects, which can be located in free space or in waveguide junctions. In our approach, each single object is considered separately in its local coordinate system. The proposed hybrid technique is used to determine the impedance matrix, which defines a relation between electric and magnetic tangential field components on the artificial cylindrical surface surrounding analyzed object. Because the impedance matrix of each scatterer is known, the analytical iterative scattering procedure can be applied to determine the scattering parameters of arbitrary configuration of objects [5, 11]. Presented analysis is an extension of our work presented in Refs. 12–14, where the cylindrical objects with arbitrary cross-section were analyzed. In this study, the method is modified to the analysis of objects with axial symmetry. To check the validity of the proposed approach, the
Closed-form design method for tight parallel-coupled microstrip coupler with ultra-wideband performance and practical dimensions

A.M. Abbosh

A closed-form method for designing a microstrip coupler that has a simple planar structure, tight coupling, practical dimensions, and ultra-wideband performance is presented. According to the proposed method, the coupled microstrip structure is divided into three sections. A theoretical model based on the even-odd mode analysis of four-port networks is derived and used to find the optimum length and coupling factor for each of those sections to get an ultra-wideband performance. According to the model, the central section is designed to have a tight coupling by utilising a slotted ground plane and one lumped capacitor, whereas the two side sections have loose coupling. The presented method is validated by building a compact 3 dB coupler that has more than 116% fractional bandwidth.

Introduction: Tight microstrip couplers are key elements in many circuits, such as balanced mixers, amplifiers, and beamforming networks of antenna arrays. The parallel-coupled microstrip lines can easily provide loose coupling between 10 and 20 dB [1]. However, the conventional structure of parallel-coupled microstrip couplers cannot be used to achieve a tight coupling, such as 3 dB, owing to the need for an impractically narrow spacing between the two coupled lines [2].

Several methods have been proposed to enhance the coupling of parallel-coupled lines across a wideband [2–6]. A slotted ground plane is used in [2, 4] underneath the coupled lines to ease the narrow gap requirement and to realise up to 70% fractional bandwidth. In [3], a cavity under the coupled lines is used to enhance the inductive coupling, and thus, to obtain a tight coupling across a band from 1 to 3 GHz. However, the requirement for a bulk cavity underneath the coupled structure complicates the design. In another approach [5], more than 70 short sections of coupled coplanar waveguides are connected together and overlaid with dielectric and conductive layers that are supported by air-bridges. Although the design is able to achieve 3 ± 1 dB coupling over 70% fractional bandwidth, the utilised structure is too complicated. In [6], vertical metallic plates are connected with parallel-coupled microstrip lines. A material with high dielectric constant is inserted between the two vertical plates to enhance the odd-mode capacitor of the structure. The measured performance of the proposed structure shows around 65% fractional bandwidth. The main drawback of the utilised design is its three-dimensional configuration.

In this Letter, a method that enables the building of tight parallel-coupled microstrip couplers with practical dimensions, ultra-wideband (UWB) performance, and a simple planar structure is proposed. According to the presented method, the coupled structure is divided into three short sections. A theoretical model is used to find the optimum lengths of the sections and the coupling factor for each of those sections so that an ultra-wideband performance can be realised using a compact and easy-to-manufacture planar structure.

Theory and design: In the proposed method, which can be applied to any coupled configuration, the coupled structure is divided into three sections. For a symmetrical configuration, the two side sections have the same coupling factor (k1), whereas the central section has a coupling factor of k2 and length l2. Using the even-odd mode analysis approach for four-port devices [1], it is possible to show that the normalised direct (S11) and coupled (S13) output powers of the three-section coupled structure are given as

\[ S_{31} = \frac{\beta^2 \beta_2}{(1 - \alpha_1 \alpha_2^2) - \alpha^2 \beta_2^2} \]  
\[ S_{31} = \alpha_1 + \frac{\beta^2}{(1 - \alpha_1 \alpha_2^2) - \alpha^2 \beta_2^2} \]  
\[ \alpha_1 = \frac{j \sqrt{2 \rho (2 \pi l_1 / \lambda_0)}}{1 - k^2 \cos(2 \pi l_1 / \lambda_0) + j \sin(2 \pi l_1 / \lambda_0)} \]  
\[ \beta_2 = \frac{j \sqrt{2 \rho (2 \pi l_2 / \lambda_0)}}{1 - k^2 \cos(2 \pi l_2 / \lambda_0) + j \sin(2 \pi l_2 / \lambda_0)} \]

where \( k \) and \( l \) are the coupling factor and the length, respectively, for the \( i \)th coupled section, and \( \lambda_0 \) is the effective wavelength.

The iterative solution of (1)–(4) shows that it is possible to obtain a 3 ± 1 dB overall coupling across the band from 3.1 to 10.6 GHz when the length and the coupling factor are: \( l_1 = 0.08 \lambda_0 \), \( k_1 = 0.4 \) for the two side sections, and \( l_2 = 0.15 \lambda_0 \), \( k_2 = 0.79 \) for the central section. \( \lambda_0 \) is the effective wavelength calculated at the centre frequency of operation, which is equal to 6.85 GHz for the UWB (3.1 to 10.6 GHz) operation. The overall length of the coupled structure is equal to 0.31 \( \lambda_0 \).

It is easy to achieve the required loose coupling of 0.4 (equivalent to ~8 dB) at the two side sections using the conventional parallel-coupled microstrip lines. However, the tight coupling of 0.79 (equivalent to ~2 dB) at the central section cannot be realised across an UWB using the conventional structure. Thus, two techniques are employed: a slotted ground plane is used underneath the central section in order to decrease the even-mode capacitor, and thus, to increase the even-mode impedance [2, 4], and a lumped capacitor is connected between the two coupled lines at the centre of that section to increase its odd-mode capacitor, and thus, to decrease its odd-mode impedance [5, 6]. The proposed method is implemented using parallel-coupled microstrip lines in the manner shown in Fig. 1. The coupled lines and the lumped capacitor are at the top side, whereas the slotted ground is at the bottom side.

![Fig. 1 Proposed parallel-coupled microstrip coupler](image)

One of the main differences between the design technique proposed in this Letter and other methods [2–6] is that it is required here to obtain a tight coupling at a short (\(< \lambda_0 / 4\)) coupled section, which can be implemented easily without complicating the structure. In the other methods, the tight coupling is to be achieved across the whole structure resulting in complicated configurations [3, 5, 6] or limited bandwidths [2, 4].

To ease the manufacturing requirements of the coupler using the printed circuit board technology, the spacing \( s \) between the coupled lines is set at 0.15 mm. With this value of \( s \) and for a certain substrate, the width of the side sections (\( w_1 \)) to achieve the required coupling of 0.4 are found using the equations presented in [2]. Concerning the central section, the analysis for parallel-coupled lines with slotted ground [2] is employed after including the effect of the additional lumped capacitor.

To that end, the cross-sectional view of the central section shown in Fig. 2 is used to find the equivalent even- (\( C_e \)) and odd-mode (\( C_o \)) capacitors. For the even-mode, the line of symmetry shown in Fig. 2 behaves as a magnetic wall (H-wall), whereas it behaves as an electric wall (E-wall) in the odd-mode. The even- and odd-mode capacitors are, thus, equal to

\[ C_e = C_{G1} ; \quad C_o = C_{G2} + 2C_a + 2C_{G2} \]

where \( C_e \): the capacitor per unit length between each of the coupled lines and the ground, \( C_o \) and \( C_a \) the mutual capacitor between the two coupled lines in free space and dielectric, respectively, and \( C_{G2} \) the added
The impedance at, say the \( i \)th mode, \( Z_{oi} \) can be calculated from the \( i \)th mode capacitor \( C_i \) \[1\]

\[
Z_{oi} = \frac{1}{c\sqrt{C_{io}C_i}}
\]  

\( C_i \) the \( i \)th mode capacitor after replacing the substrate with air, and \( c \) the speed of light in free space.

Using the well-known equations that relate the coupling factor with the mode impedances \[1\], it is possible to show that the even- and odd-mode impedances of the central section should equal to \( 17.1 \) and 146 \( \Omega \), respectively, so that the coupling of this section is 0.79. Using these values for the mode impedances, the dimensions of the central section can be found from \( (5), (6) \) and the relation between the capacitors \( (C_{io}, C_i, C_d) \) and the lines’ dimensions as derived in \[2\].

\[ \text{Results and discussion:} \]

To test the accuracy of the proposed design method, a 3 dB directional coupler was designed and fabricated using the substrate RT6010 (dielectric constant = 10.2, thickness = 0.635 mm). According to the presented design method and with the help of the optimisation capability of the software CST Microwave Studio, the following design values were obtained assuming that \( s \) is fixed at 0.15 mm: \( w_1 = w_2 = 0.55 \) mm, \( l_1 = 1.3 \) mm, \( l_2 = 2.2 \) mm, \( C_0 = 0.22 \) pF/mm, and width of the ground slot = 5.1 mm. Practically, the value of the utilised lumped capacitor, which is a broadband microwave chip capacitor, equals to \( C_d = 5 \) pF. As shown in Fig. 1, a tapered microstrip is used to connect the coupled structure with the input/output ports. The overall dimensions of the developed coupler is \( 2 \times 2 \) cm.

The performance of the designed device was verified via simulations (CST Microwave Studio) and measurements. According to the results depicted in Fig. 3, the value of the coupling factor is 3 dB ±1 dB across the band from 3.1 to 13.1 GHz in the simulations and from 3.3 to 12.5 GHz in the measured results. Thus, the achieved fractional bandwidth is more than 116%. As revealed in Fig. 3, the return loss at any of the ports (due to symmetry, the ports have the same return loss performance) and the isolation between the two output ports are better than 20 dB in the simulated results and 17 dB in the measured results across the whole investigated band from 2 to 14 GHz. Concerning phase performance, the designed coupler is quadrature as the measured phase difference between the two output ports is 90° ± 3° across the band from 2 to 14 GHz as indicated in Fig. 3. The simulated and measured performances of the coupler agree well with each other as shown in Fig. 3. The slight difference, especially at the upper end of the investigated band, is thought to be due to the utilised lumped capacitor, which is not expected to have the same value across the whole band.

\[ \text{Conclusion:} \]

A closed-form design procedure for microstrip couplers that have tight coupling, practical dimensions, simple planar structure, and ultra-wideband performance is presented. According to the proposed method, the coupled structure is divided into three sections. A theoretical model to find the optimum values for the length and coupling factor for each of those sections has been derived. The simulated and measured results for a 3 dB coupler designed using the proposed method prove its validity.

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One or more of the Figures in this Letter are available in colour online.
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References

Figure 10 shows the radiation patterns of manufactured antennas. The measured results confirm that backside radiation level of the proposed antenna maintains a value of $-16$ dBi. Therefore, the proposed antenna has a front to back ratio of 21 dB. The proposed antenna maintains the directional radiation property, which is the main advantage for a patch antenna.

Figure 11 shows the gain of both the original antenna and the SRR-loaded antenna. According to the results, the designed antenna achieves a maximum gain of $-4.84$ dBi which is about 0.2 dBi below that of the original antenna.

4. CONCLUSIONS
This article proposes the design of a compact microstrip patch antenna. To achieve size reduction, a modified substrate with SRR (split-ring resonators) structures is utilized. The resonant frequency of the fabricated SRR-loaded patch antenna shifts from 6.6 GHz (i.e., the resonant frequency of the original patch antenna) to 4.67 GHz, achieving a $29.3\%$ antenna size reduction. The gain of the proposed antenna reaches 4.84 dBi. Moreover, the directional radiation property of a patch antenna maintains. These properties make the proposed design useful for various applications.

ACKNOWLEDGMENT
This work was supported by the IT R&D program of MKE/KCC/KEI (K1002071), for the Study of technologies for improving RF spectrum characteristics by using the meta-electromagnetic structure.

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The parallel-coupled microstrip lines can easily provide loose coupling in the range from 10 to 40 dB and reasonable directivity, especially if the compensation techniques are used [4–10]. However, the conventional parallel-coupled microstrip couplers cannot be used to achieve a tight coupling, such as 3 dB, due to the need for an impractically narrow spacing between the two-coupled lines [11].

Several methods have been proposed to enhance the coupling of parallel-coupled lines [12–16]. In Refs. 12–14, different shapes of slotted ground planes are used underneath the coupled lines to ease the narrow gap requirement. However, these structures impose limitations on the packaging and integration of the coupler with other devices since a reasonable space is required underneath the slotted ground plane to avoid any disturbance to its performance. In another approach [15], more than seventy short sections of coupled coplanar waveguides are connected together and overlaid with dielectric and conductive layers that are supported by air-bridges [15]. The design is able to achieve 3 dB ± 1 dB coupling over 70% fractional bandwidth. However, the utilized structure is complicated. In Ref. 16, vertical metallic plates are connected with the parallel-coupled microstrip lines. A high dielectric constant material is inserted between the two vertical plates to enhance the odd-mode capacitor of the structure. The measured performance of the proposed structure shows around 65% fractional bandwidth. The main drawback of the utilized design is its three-dimensional configuration.

In this article, a closed-form method is presented to enable building tight parallel-coupled microstrip couplers with a practical spacing between the coupled lines and simple structure. The presented method is based on the quasi-static even-odd mode technique. To relax the requirement on the dimensions of the coupled structure, three-lumped capacitor are connected symmetrically between the coupled lines. A complete design procedure is presented and validated by building two 3 dB couplers using two different substrates.

2. THEORY AND DESIGN

The parallel-coupled microstrip coupler (Fig. 1) can be analyzed using the quasi-static even-odd mode approach. Under this approach, the even- \( Z_{ee} \) and odd-mode \( Z_{oo} \) impedances of the lines can be defined using, respectively, the even- \( C_e \) and odd-mode \( C_o \) capacitances per unit length according to the following equations [1, 2, 11]

\[
Z_{ee} = \sqrt{\frac{\varepsilon_0}{\epsilon C_e}} \quad (1)
\]

\[
C_e = C_\text{g} \quad (2)
\]

\[
Z_{oo} = \sqrt{\frac{\varepsilon_0}{\epsilon C_o}} \quad (3)
\]

\[
C_o = C_\text{g} + 2C_{ad} \quad (4)
\]

In Eqs. (1) and (3), \( \varepsilon_e \) and \( \varepsilon_o \) are the effective even- and odd-mode dielectric constant of the coupled structure, respectively, and \( \epsilon \) is speed of light in free space. The different capacitances used in Eqs. (2) and (4) are shown in Figure 2; \( C_\text{g} \) is the effective capacitance per unit length between any of the coupled lines and the ground, whereas \( C_{ad} \) is the capacitance per unit length between the coupled lines. As shown in Figure 2, electric (\( E \)) and magnetic (\( H \)) walls are used when calculating the value of capacitors via, for example, the conformal mapping technique.

The coupling factor \( K \) is given as

\[
K = \frac{Z_{ee} - Z_{oo}}{Z_{ee} + Z_{oo}} \quad (5)
\]

Assuming, as a rough approximation not used in the design procedure but only for the method’s explanation, that \( \varepsilon_{ee} = \varepsilon_{oo} \), the coupling factor \( K \) can be approximated as

\[
K = \frac{C_{ad}}{(C_\text{g} + C_{ad})} \quad (6)
\]

The capacitance \( C_\text{g} \) depends mainly on the width of the coupled lines and characteristics of the substrate [1, 11]. The spacing \( s \) has negligible effect on its value. However, the capacitance \( C_{ad} \) increases as the gap narrows [11]. To achieve a tight coupling, \( C_{ad} \) should be relatively large as can be concluded from Eq. (6). For the conventional microstrip coupler of Figure 1, this condition can only be achieved with impractical value for \( s \).

The strategy proposed in this article is to introduce an additional factor in Eq. (6) to compensate for the low value of \( C_{ad} \) when a wide gap is used for practical reasons. Assume that distributed lumped capacitors are connected between the coupled lines in the manner shown in Figure 3. The even-mode
or, alternatively, in terms of the capacitances with the rough approximation that assumes \( \varepsilon_{cv} = \varepsilon_v \)

\[
K_m = \frac{(C_{ad} + C_x)}{(C_v + C_{ad} + C_x)}
\]  

(10)

Therefore, the gap between the coupled lines can be increased to a reasonable and practical value as \( C_v \) can be used to compensate for the reduction in \( C_{ad} \) and the tight coupling can still be achieved.

For a perfect matching between the coupled section and the input and output ports, which have characteristic impedance \( Z_0 = 50 \, \Omega \), \( Z_{nom} \) and \( Z_{oom} \) should relate to \( Z_0 \) according to the following equation

\[
Z_m = \sqrt{Z_{nom}Z_{oom}}
\]  

(11)

In summary, the design approach is

1. Depending on the available manufacturing tools, choose a practical value for the gap \( s \).
2. Depending on the required coupling factor \( K_m \), calculate the required even- and odd-mode impedances of the modified structure \( Z_{oom} \) and \( Z_{oom} \) using Eqs. (9) and (11).
3. Using \( Z_{oom} \), \( s \), and characteristics of the utilized substrate (the dielectric constant \( \varepsilon_v \) and the thickness \( h \)), calculate the required width of the coupled lines \( w_c \), \( t_{ee} \), and \( C_v \) using the equations presented in Ref. 11 and Eq. (1).
4. Using \( w_c \), \( s \), \( \varepsilon_v \), and \( h \), calculate \( \varepsilon_{cv} \), \( C_{ad} \), and \( C_x \) of the coupled structure without including effect of the added capacitors using the equations presented in Ref. 11.
5. Calculate the required \( C_x \) using Eq. (8). This capacitor is the per unit length capacitor distributed uniformly as shown in Figure 3. The actual capacitor needed to complete the design is \( C_x/l \), where \( l \) is the length of the coupled structure and it is taken as quarter of the effective wavelength calculated at the center frequency \( f_c \), i.e.,

\[
l = \frac{c}{2f_c\sqrt{\varepsilon_{ee} + \varepsilon_{cv}/2}}
\]

Practically, it is difficult to connect a distributed rack of capacitors as suggested in Figure 3. The reasonable alternative is to divide the coupled structure into, say, three sections and connect a capacitor at the center of each section. Thus, three capacitors with a total value of \( C_x \) are needed. It is noted via parametric analysis that representing \( C_x/l \) by one capacitor connected at the center of the structure can achieve the required coupling factor at the center frequency, but it is not a good option for broadband performance.

The remaining question is whether the presented method can be used to achieve a tight coupling for any value of the gap \( s \) between the coupled lines. Following the procedure of the proposed method, it is possible to show that increasing \( s \) requires an increase in \( C_v \). A parametric analysis using CST Microwave Studio confirms this conclusion. However, the same parametric analysis shows that with relatively wider gaps and larger capacitors, the bandwidth narrows. Thus, the minimum achievable gap with the available manufacturing tools is to be used for broadband performance.

3. RESULTS AND DISCUSSION

To test the accuracy of the proposed design method, 3 dB directional couplers were designed and fabricated using RT6010 (\( t_s = 10.2 \) and \( h = 1.27 \text{ mm} \); coupler #1), and RO4003 (\( t_s = 3.38 \) and \( h = 0.813 \text{ mm} \); coupler #2) as the substrates. Using the proposed design steps, the following design values were obtained assuming \( f_c = 7 \text{ GHz} \) and \( s = 0.1 \text{ mm} \). For coupler #1, \( w_c = 0.36 \text{ mm} \), \( l = 4.3 \text{ mm} \), and \( C_x/l = 0.33 \text{ pF} \). For coupler #2, \( w_c = 0.63 \text{ mm} \), \( l = 6.7 \text{ mm} \), and \( C_x/l = 0.48 \text{ pF} \). The suitable broadband microwave capacitors that are available to the author have the values of 0.1 and 0.2 pF from Johanson Technology [17]. Thus, three 0.1 pF capacitors are used for the coupler #1, whereas a capacitor of 0.1 pF at the center and two capacitors of 0.2 pF at the sides are used for coupler #2. To compensate for the slight difference between the calculated and available capacitors and to get the best possible performance, the design parameters were optimized using HFSS. The final values are:

- For coupler #1 \( w_c = 0.35 \text{ mm} \) and \( l = 3.9 \text{ mm} \) and for coupler #2 \( w_c = 0.61 \text{ mm} \) and \( l = 6.3 \text{ mm} \).

Comparing the initial design values with the optimized values shows that they are very close to each other, and thus, validates the presented design method. A photo of one of the developed devices is depicted in Figure 5.

![Figure 3](image_url)  
**Figure 3** The modified parallel-coupled microstrip lines. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

![Figure 4](image_url)  
**Figure 4** The odd-mode capacitances (per unit length) of the modified coupler. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
To demonstrate the effectiveness of the proposed design approach in relaxing the gap requirement, it is possible to show that the gap $s$ needed for a 3 dB microstrip coupler without capacitors (Fig. 1) is 0.005 mm for coupler #1 and 0.0003 mm for coupler #2. Those values are clearly impractical for the PCB technology.

The performance of the designed devices was verified using the software CST Microwave Studio and then measured using a vector network analyzer. The simulated and measured performance of the coupler #1 and #2 agree well with each other as shown in Figures 6 and 7, respectively. The value of the coupling factor for coupler #1 is 3 dB $\pm$ 1 dB across the band from 3.8 to 9.3 GHz in the simulations and from 4 to 9 GHz in the measured results. Concerning coupler #2, the coupling factor is 3 dB $\pm$ 1 dB across the band 3.9–9.6 GHz, which is equivalent to 84% fractional bandwidth, according to the simulated and measured results. Figures 6 and 7 also show the return loss at any of the three ports (due to symmetry $S_{11} = S_{22} = S_{33}$), and the isolation between the two output ports. Those values are better than 15 dB for coupler #1 and 13 dB for coupler #2 across the abovementioned bands. Concerning the phase performance of the developed couplers, Figures 6 and 7 reveal that the two devices are quadrature couplers as the phase difference between the two output ports is 90$^\circ$ $\pm$ 1$^\circ$ in the simulations and 90$^\circ$ $\pm$ 2$^\circ$ in the measurements for the two couplers across the band 3–10 GHz.

Some applications may need as high as 30 dB isolation between the output ports. For those applications, any of the compensation techniques [4–10] can be utilized in the proposed coupler. However, the target in this article is to present a closed-form method to build a tight parallel-coupled microstrip coupler with broadband performance and reasonable isolation using a simple circuit that does not need any vias and/or slotted ground.

4. CONCLUSION

A complete design procedure for microstrip couplers that have tight coupling, practical dimensions, simple structure, and broadband performance has been presented. To enable the use of easy-to-manufacture dimensions for the coupled structure, the design method assumes the use of three-lumped capacitor that are connected between the centers of the coupled lines. The simulated and measured results for two 3 dB couplers designed using two different substrates validate the presented method.

REFERENCES

1. INTRODUCTION

In recent years, vehicular communications suffered a great development in response to the various traffic problems that affect the users in their daily life, such as accidents, congestions, pollution, etc. These communications are based on wireless communication technology, known as dedicated short range communications (DSRC) [1]. In addition to safety, efficiency, and navigation aids, this technology also enables other more conventional applications, such as electronic toll collection, automatic payment in parks and gas stations, etc.

2. RELATED WORK AND CONTRIBUTION

Circular polarization is used to reduce the interference caused by the reflected waves, and can be achieved through changes in the physical structure of the elements of the array or by changes to their feeding method. The simplest way to build a circularly polarized microstrip antenna is by changing its physical structure. However, by just doing so, it results in a very narrow bandwidth antenna. To obtain a circularly polarized antenna another technique is the sequential rotation feeding in phase and outside this volume should not exceed 30° in the horizontal plane so as to avoid interference in communication of other OBUs on vehicles that circulate in adjacent lanes. The equivalent isotropic radiated power (EIRP) is limited to +33 dBm up to an angle of 70° to the vertical, and outside this volume should not exceed +18 dBm. To feed the antenna at the maximum allowed power level, the EIRP of the side lobes should be 15 dB below the EIRP at boresight. Polarization should be left circular with a cross-polarization rejection of over 15 dB in boresight and greater than 10 dB in the directions where the gain drops 3 dB.

In this band of frequencies, microstrip antennas have been increasingly used as they have important advantages such as its low profile, weight and cost, high efficiency and ease of manufacture, which make them suitable for road communications. However, in microstrip antennas, the center frequency may undergo a shift due to small deviations in their dimensions or in dielectric constant of the substrate. A wide bandwidth of the antenna is important because it increases the tolerance to those deviations, which is needed to achieve high yields in mass production.

Moreover, the DSRC technology for intelligent transportation systems in Europe operates in other frequency band ranging from 5.855 to 5.925 GHz, for safety, nonsafety, and other future applications. An antenna that operates over both DSRC bands would be advantageous because the same antenna could be used in various applications.
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WIDEBAND PERFORMANCE OF 3 dB MICROSTRIP-SLOT COUPLER USING DIFFERENT SUBSTRATES

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ABSTRACT: This article presents the design of an elliptically shaped microstrip-slot coupler using different thickness and permittivity substrates that is aimed for operation in 3.0–11.0 GHz band. CST Microwave Studio simulator is applied to its design and performance assessment. Design with the best performance is fabricated to validate the investigation. It is shown that the wideband operation of this type of coupler is predominantly dependent on the substrate’s thickness and to a lesser degree on the substrate’s permittivity. The degradation in performance is observed for an increased substrate thickness and is explained by the presence of a higher order mode. When the substrate thickness is fixed to a small value and the cutoff frequency of the higher mode is outside the investigated band, there is an optimal permittivity, which offers the best coupler’s performance in terms of coupling coefficient, return loss, isolation, and phase difference between the two output ports. In the present investigation, it is found that the best operation is obtained when the substrate’s relative permittivity is 4.5 and the substrate thickness of 0.508 mm. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 53:1618–1624, 2011; View this article online at wileyonlinelibrary.com. DOI 10.1002/mp.26082

Key words: directional couplers; microstrip coupler; coupler design

1. INTRODUCTION

It has been demonstrated recently that the application of the microstrip-slot technique using a double-layer dielectric substrate with a common ground slot can lead to the design of a fine quality 3-dB directional coupler operating over an ultra-wide band (UWB) [1]. The design shown in [1] follows an initial idea introduced in [2] where two microstrip lines coupled by a rectangular slot in a common ground plane were used to form a coupler. A simple manufacturing process and a good performance of this coupler triggered further investigations by other researchers. For example, in [3] the coplanar waveguide technique was applied to obtain a wideband 3-dB coupler. In turn, in [3, 4], a hexagonal-shaped coupler in multilayer microstrip technology was designed. In [5, 6, 7], multiple-section configurations were explored to further enhance the operational bandwidth of the coupler in [1]. In all of these quoted works, a wideband operation with respect to the coupling coefficient, return loss and isolation has been demonstrated. However, none of
these works posed the question of whether a similar superior performance can be achieved if the design uses substrates of different thickness or permittivity.

This article responds to this question and shows that the best performance can only be obtained for substrates with properly selected permittivity and thickness. The investigations focus on the elliptical coupler configuration. This choice is motivated by the work in [8] where it has been pointed out that the elliptical shape leads to the best performance of the 3-dB microstrip-slot coupler.

The undertaken investigations start with the initial 3-dB elliptical microstrip-slot coupler developed on the 0.508-mm thick RO4003C substrate as reported in [1]. Then, the investigation continues by designing the coupler on a fixed thickness substrate of 0.508 mm with varied permittivity. Next, the design assumes a few substrates having the same permittivity but different thickness. Because the works presented in [1, 5, 6, 8] have already confirmed the validity of the full-wave EM simulation approach, the study are conducted using CST Microwave Studio. Only the design with the optimum performance is fabricated and validated experimentally.

2. DESIGN USING DIFFERENT SUBSTRATES

The commercially available substrates chosen for the present investigations are shown in Table 1. The effect of the choice of these substrates on the performance of the 3-dB microstrip-slot coupler is investigated in two stages. In the first stage, the substrate thickness is fixed at 0.508 mm, but its relative permittivity is varied. The variations are accomplished using six substrates with the permittivity in the range of 1 to 7. To determine the optimum thickness of the optimum permittivity, the couplers are designed using few different substrates having four different thicknesses in the second stage. The first two substrates are chosen based on the optimum performance investigated in the first stage, whereas the third substrate is chosen to observe the performance of high permittivity substrate, which is not investigated in the first stage due to unavailability of 0.508 mm thickness in the market.

The configuration of the investigated coupler adopted from [1] is shown in Figure 1. The coupler is constructed using three conductor layers interleaved by two dielectric layers. A common ground is present between the two dielectrics. Ports 1 and 2 are on the top layer, whereas Ports 3 and 4 are on the bottom dielectric layer. The 3-dB quadrature of signals at Ports 2 and 3 is achieved by controlling the width of elliptical patches and ground slot. The initial dimension of $D_1$, $D_2$, $D_3$, and $w_m$ for the investigated substrates of Table 1 are obtained by following

<table>
<thead>
<tr>
<th>Substrate</th>
<th>Relative Permittivity ($\varepsilon_r$)</th>
<th>Thickness $h$ (mm)</th>
<th>$D_1$ (mm)</th>
<th>$D_2$ (mm)</th>
<th>$D_3$ (mm)</th>
<th>$w_m$ (mm)</th>
<th>Board's Dimension($x$ mm * $y$ mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>RT/durid 5870LZ</td>
<td>1.96</td>
<td>0.508</td>
<td>6.5</td>
<td>8.7</td>
<td>8.4</td>
<td>1.68</td>
<td>38 x 18</td>
</tr>
<tr>
<td>RT/durid 5870</td>
<td>2.33</td>
<td>0.508</td>
<td>6.0</td>
<td>8.7</td>
<td>8.2</td>
<td>1.51</td>
<td>36 x 16</td>
</tr>
<tr>
<td>RO3003</td>
<td>3.00</td>
<td>0.508</td>
<td>5.0</td>
<td>8.0</td>
<td>7.7</td>
<td>1.28</td>
<td>34 x 15</td>
</tr>
<tr>
<td>RO4003C</td>
<td>3.38</td>
<td>0.508</td>
<td>4.8</td>
<td>7.4</td>
<td>7.3</td>
<td>1.18</td>
<td>28 x 12</td>
</tr>
<tr>
<td>TMM4</td>
<td>4.50</td>
<td>0.508</td>
<td>4.2</td>
<td>6.8</td>
<td>6.7</td>
<td>0.96</td>
<td>24 x 10</td>
</tr>
<tr>
<td>RO4360</td>
<td>6.15</td>
<td>0.508</td>
<td>3.4</td>
<td>6.7</td>
<td>6.1</td>
<td>0.75</td>
<td>20 x 10</td>
</tr>
<tr>
<td>RO4003C</td>
<td>3.38</td>
<td>0.203</td>
<td>4.8</td>
<td>7.4</td>
<td>7.3</td>
<td>1.18</td>
<td>34 x 15</td>
</tr>
<tr>
<td>TMM4</td>
<td>6.15</td>
<td>0.127</td>
<td>4.8</td>
<td>7.4</td>
<td>7.3</td>
<td>1.18</td>
<td>28 x 12</td>
</tr>
<tr>
<td>RT/durid 6010LM</td>
<td>10.20</td>
<td>0.127</td>
<td>4.8</td>
<td>7.4</td>
<td>7.3</td>
<td>1.18</td>
<td>28 x 12</td>
</tr>
</tbody>
</table>

Figure 1 CST layout of investigated coupler. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
the guidelines described in [1]. The coupler is then simulated using CST Microwave Studio and the final design dimensions (for the coupling of 3 dB, return loss, and isolation better than 15 dB, phase imbalance less than 5°) are shown in Table 2.

It can be seen from Table 2 that when \( h \) is increased, the dimensions for each physical parameter are also increased, resulting in larger board dimensions for a larger substrate thickness. In contrast, the use of substrates with a larger relative permittivity \( e_r \) results in smaller board dimensions.

3. RESULTS

The operation of the designed couplers is investigated in 3–11 GHz frequency band. Figure 2 shows the simulation results for the 3-dB microstrip-slot coupler designed using six substrates having different permittivity but the same thickness of 0.508 mm.

Figures 2(a)–2(c) show, respectively, the results for the coupling coefficient, return loss, and isolation, whereas Figure 2(d) shows the result for the phase difference between the output ports. The results presented in Figure 2(a) reveal that for all the six chosen values (1.96, 2.33, 3, 3.38, 4.5, and 6.15) of relative permittivity \( e_r \), the simulated coupling coefficient is 3 ± 2 dB in the entire 3–11 GHz band. The return loss and isolation are better than 18 dB in the same band, and the output port phase difference deviation from the nominal value of 90° is about ±3°.

As observed in Figure 2(a), the coupling performance for each \( e_r \) is close to each other. The best return loss and isolation can be seen for \( e_r = 4.5 \) followed by \( e_r = 3.38 \). Their values for \( e_r = 4.5 \) are not less than 27 dB, while for \( e_r = 3.38 \) are not less than 23 dB across UWB. The output phase imbalance for both permittivity is only 0.3° and 0.6°, respectively, which represents an excellent result. By considering coupling, return loss, isolation, and phase difference of thickness \( h = 0.508 \) mm, substrate TMM4 provides the best performance followed by the substrate RO4003C.

Figure 3 shows the simulated amplitude of the scattering parameters and phase difference between the two output ports for the designed 3-dB couplers using four different values of RO4003C substrate thickness \( h \).

As observed in Figure 3(a), the coupling coefficient is 3 ± 2 dB for \( h = 0.203, h = 0.508, \) and \( h = 0.813 \) across UWB spanning from 3.1 to 10.6 GHz. However, the mid-band coupling value for \( h = 0.813 \) is 3 ± 1.2 dB, which is slightly degraded compared to \( h = 0.203 \) and \( h = 0.508 \). In turn, for \( h = 1.524 \), the operating bandwidth is drastically reduced to 3–6 GHz band. However, in this limited band the coupling imbalance is improved and is less than 1 dB.

Figures 3(b)–3(d) show performance of the coupler for the four selected values of substrate thickness for the remaining parameters that include return loss, isolation, and phase difference between the output ports. The return loss and isolation, presented in Figures 3(b) and 3(c), are better than 15 dB across UWB band for all thicknesses except for \( h = 1.524 \). For this thickness, the return loss and isolation bandwidth is limited to 3–9.2 and 3–6.3 GHz, respectively. Concerning the phase difference between two output ports, the phase imbalance is less than 5° for the 3–11 GHz band for \( h = 0.813 \) or smaller. For \( h = 1.524 \), the upper operating frequency for the phase difference of 90° ± 5° is limited to 7.5 GHz. Also, most of the values are between 86° to 88° instead of 90°. By taking into account all four parameters (coupling, return loss, isolation, and phase difference), it can stated that the coupler designed on RO4003C substrate with \( h = 0.813 \) or smaller features UWB.
Figure 3  Simulated performance using four different values of RO403C substrate thickness $h$ (a) coupling coefficient, (b) return loss, (c) isolation, and (d) phase difference between the two output ports. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 4  Simulated performance using four different values of TMM4 substrate thickness $h$ (a) coupling coefficient, (b) return loss, (c) isolation, and (d) phase difference between the two output ports. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
characteristics across 3.1–10.6 GHz while, for $h = 1.524$, the acceptable bandwidth is much lower than for the thinner substrate counterparts and is confined to 3–6 GHz.

Figures 4 and 5 shows the simulated amplitude of the scattering parameters and phase difference between the two output ports for the designed 3-dB couplers using TMM4 and RT6010 substrate for four different cases of thickness $h$.

In the case of TMM4 substrate, considering a coupling coefficient of $3 \pm 2$ dB, the coupling performance is within the UWB band for $h = 0.762$ mm and lower, whereas the coupling bandwidth is limited to 3–6 GHz when $h = 1.524$ mm. From Figures 4(b) and 4(c), the return loss and isolation are better than 15 dB across UWB band for all thicknesses except when $h = 1.524$ mm. The isolation bandwidth is limited to 3–7.5 GHz when considering isolation better than 10 dB. In Figure 4(d), the best performance of the phase difference between the output ports can be seen when $h = 0.508$ mm. By assuming the following criteria (coupling coefficient of $3 \pm 1$ dB, return loss, and isolation better than 25 dB), the optimum design on TMM4 can be obtained when $h = 0.508$ mm.

In the case of RT6010 substrate, considering a coupling coefficient of $3 \pm 2$ dB, return loss, and isolation better than 10 dB and phase imbalance of 5°, the coupler’s bandwidth is 3.5–8.5 GHz for $h = 0.127$ mm, 3.5–11 GHz for $h = 0.254$ mm, 3.5–10 GHz for $h = 0.635$ mm, and 4.5–6 GHz for $h = 1.27$ mm.

According to Figures 3–5, it can be seen that when $h = 1.27$ mm and $h = 1.524$ mm, the coupler fails to operate across the UWB frequency band in terms of the coupling coefficient, return loss, isolation, and phase imbalance. Based on these simulations, assuming coupling $3 \pm 1$ dB, return loss and isolation better than 25 dB, the optimum design of the substrate can be obtained when dielectric constant is 4.5 and thickness of 0.508 mm. This is the case of TMM4 substrate.

A prototype of an elliptical coupler manufactured on TMM4 substrate is shown in Figure 6. This coupler was experimentally tested and results were compared with simulations. The result of measurements concerning amplitude of the s-parameter is shown in Figure 7 alongside the simulated result. These results show a
3-dB quadrature and UWB performance, and indicate a good agreement between the simulated and measured values. The observed discrepancies can be explained by the use of coaxial to microstrip transitions in the measurement system, which were not included in simulations. Also, tiny air gaps in the developed prototype could be the factor behind these differences. Nevertheless, the experimental data confirms an excellent performance of this coupler in the 3–11 GHz band.

4. DISCUSSION

The worsening performance of some of the designed couplers can be explained by the higher order mode theory described in [9]. One has to note that to meet the mid-band 3-dB coupling value, the odd \( Z_{0o} \) and even mode \( Z_{0e} \) characteristic impedances have to take the values \( Z_{0o} = 20.7 \Omega \) and \( Z_{0e} = 120.5 \Omega \), respectively [1]. Note that the equivalent two-ports require the slot to be closed by the magnetic conductor for the even mode, whereas the electric conductor is required for closing the slot for the odd mode. As a result, the odd mode impedance \( Z_{0o} \) of 20.7 \( \Omega \) implies that the width of the microstrip line forming the microstrip-slot coupler’s plate has to be wide. This wide microstrip is prone to launching a higher order mode. This effect becomes more pronounced when the substrate’s thickness is increased. In this case, the microstrip line width has to also be increased to maintain the low value of odd-mode characteristic impedance. A higher order mode, which can be excited in such a case travels with a different phase velocity to that of the fundamental mode. The cutoff frequency, \( f_c \), for the higher order mode is given approximately by [9]:

\[
f_c = \frac{300}{\sqrt{\varepsilon_r (2W + 0.8h)}}
\]

where \( f_c \) is in specified in GHz and \( W \) and \( h \) are in mm. It is evident from Eq. (1) that excitation of higher order mode is more obvious for thicker substrates (large \( h \)). When \( h \) is increased, the cutoff frequency of the higher mode is decreased and consequently the coupler’s bandwidth becomes reduced. This can be verified by substituting the parameter \( \varepsilon_r, W \), and \( h \) into expression (1), assuming the line’s width \( W \) giving the odd mode impedance \( Z_{0o} \) of 20.7 \( \Omega \) is calculated using the conventional microstrip design formula.

The relationship can be clearly seen in the graph form as presented in Figure 8. It can be deduced from Figure 8 that \( f_c \) decreases as the substrate height increases while \( f_c \) does not change much for different values of \( \varepsilon_r \). The comparison between the curves presented in Figure 8 and the coupler performance curves shown in Figures 3–5, provides evidence that the coupler developed on a substrate with thickness higher than \( h = 0.813 \) mm fails to achieve the UWB performance because of the existence of a higher order mode at the upper end of UWB. This higher order mode starts to propagate at frequencies above 6.6 GHz. In turn, for substrate thickness \( h = 0.813 \) mm and lower, changes in \( \varepsilon_r \) do not lead to launching the higher order mode. This is confirmed by the results for the higher mode cutoff frequency presented in Figure 8, which show that for \( h = 0.813 \) mm and lower, \( f_c \) is greater than 12 GHz for \( \varepsilon_r \) from 1 to 11. Figure 8 also shows that for any other chosen value of \( h \), the cutoff frequency of the higher order mode only slightly depends on the substrate permittivity when the odd mode of microstrip line fulfills the condition of \( Z_{0o} = 20.7 \Omega \).

5. CONCLUSIONS

In this article, investigations have been performed into wideband behavior of an elliptically shaped microstrip-slot coupler designed on substrates with different dielectric constant and thickness. The investigations have been accomplished with the use of commercially available full-wave CAD package, CST Microwave Studio. The performance parameters included coupling coefficient, return loss, isolation, and phase difference between two output ports in the frequency band of 3–11 GHz. It has been shown through the undertaken computer simulations that the wideband performance of this type of coupler is predominantly dependent on the substrate’s thickness and to a lesser degree on the substrate’s permittivity. The degradation in performance is mainly due to the presence of a higher order mode, which this coupler launches when the substrate’s thickness is increased. When the substrate thickness is fixed in such a way that the cutoff frequency of the higher mode is outside UWB, there is the optimal value of permittivity, which offers the best coupler’s performance in terms of coupling coefficient, return loss, isolation, and phase difference between two output ports. For the substrates chosen in this study, the coupler designed on the 0.508-mm thick Rogers TMM4 substrate having \( \varepsilon_r \) of 4.5 gives best performance.

Figure 7 Simulated and measured results of the fabricated TMM4 coupler. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 8 Plot of cutoff frequency, \( f_c \), versus dielectric constant, \( \varepsilon_r \), of the first higher order mode for different substrate’s thickness. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
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REFERENCES

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ENHANCEMENT OF POWER AND FREQUENCY IN PLANAR GUNN DIODES BY INTRODUCING EXTRA D̄ETAL-DOPING LAYERS
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ABSTRACT: Planar Gunn diodes operating above 100 GHz fabricated in GaAs/AlGaAs heterostructures have many advantages over conventional Gunn devices including lithographic control of operating frequency and the potential for microwave monolithic integrated circuits (MMIC) compatibility that meets the growing demand for small, cheap, and compact size. The authors present detailed studies of the improvement achieved by introducing an extra d̄etal-doping layer in the Al0.23Ga0.77As/GaAs heterostructure, together with experimental and theoretical results. The results show enhanced RF power and oscillation frequency when double d̄etal-doping technology was used. By using a two-dimensional numerical simulation tool, the conduction band profile, electron concentration in the epitaxy layers and current-voltage characteristics are investigated. Simulation results indicate that extra d̄etal-doping layers increase electron confinement in the conducting channel, therefore higher current levels are obtained. Simulated current-voltage characteristics in both cases agree well with experimental results.

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Key words: Gunn diodes; GaAs/AlGaAs heterojunctions; d̄etal-doping; numerical simulation

1. INTRODUCTION
Planar Gunn diodes have demonstrated oscillation in the fundamental transit-time mode at 108 GHz [1]. These planar Gunn devices have many advantages over conventional Gunn devices including lithographic control of operating frequency and the potential for microwave monolithic integrated circuits (MMIC) compatibility that meets the growing demand for small, cheap, and compact size. The authors present detailed studies of the improvement achieved by introducing an extra d̄etal-doping layer in the Al0.23Ga0.77As/GaAs heterostructure, together with experimental and theoretical results. The results show enhanced RF power and oscillation frequency when double d̄etal-doping technology was used. By using a two-dimensional numerical simulation tool, the conduction band profile, electron concentration in the epitaxy layers and current-voltage characteristics are investigated. Simulation results indicate that extra d̄etal-doping layers increase electron confinement in the conducting channel, therefore higher current levels are obtained. Simulated current-voltage characteristics in both cases agree well with experimental results.

2. DEVICE FABRICATION AND SIMULATION
2.1. Wafer Growth and Device Fabrication
Devices with single and double d̄etal-doping layers are shown schematically in Figure 1. The two devices have the same epitaxy layers apart from the extra d̄etal-doping layers in the Al0.23Ga0.77As layers for Figure 1 (b). The semiconductor materials were grown

Figure 1 Schematic epitaxy layers of the investigated HEMT-like planar Gunn diodes with (a) single, and (b) double d̄etal-doping layers on each side of the channel. Each d̄etal-doping layer has a sheet electron density of 8 × 1011 cm−2. The shaded areas under anodes and cathodes illustrate annealed Ohmic contact regions.
Direct quadrature phase shift keying modulation using compact wideband six-port networks

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Abstract: In this study, the operation of a compact six-port network operating as a quadrature phase shift keying (QPSK) modulator with a wide operational bandwidth is presented. The modulation has been accomplished using two types of single-layer six-port networks. The first six-port network uses a novel architecture that is newly introduced herein for use as a QPSK modulator, and employs broadband components previously developed by the same authors. The second six-port network employs the same types of broadband components in a conventional six-port architecture for comparison. The performance of the modulator using both six-port networks is tested using agilent’s advanced design system and its wideband operation is verified with measured results. It is shown that the proposed six-port modulator offers accurate QPSK symbol modulation at a symbol rate of 400 Msymbols/s across an octave bandwidth from 4.5 to 9 GHz (66% operational bandwidth).

1 Introduction

Quadrature phase shift keying (QPSK) is a digital modulation technique that modulates the angle of the cosine carrier while keeping its amplitude and frequency constant [1]. The term quadrature indicates that four output phases are possible for a single carrier frequency at a time. Since there are four possible output phases, two bits are required at the inputs, which are known as I and Q channels to produce four different output conditions; 00, 01, 10 and 11. Fig. 1 shows the block diagram of a conventional QPSK modulator [1].

As shown in Fig. 1, a conventional QPSK modulator generally uses active components and non-linear circuits, such as mixers, to modulate the signal. Owing to the use of active devices, designing a wideband modulator at high frequencies is challenging and an expensive task [2, 3]. As an alternative, a six-port network has been proposed recently to be used as a modulator [4–7], replacing high-cost mixers in direct QPSK modulators.

A six-port network is a linear network that can be formed using passive components such as power dividers and couplers. By using passive devices, a six-port network can be designed at frequencies from radio frequency (RF) [4–10] to millimetre waves [11], and has the potential to be designed for a wider operational bandwidth. In the RF and microwave bands, simple technology such as microstrip can be implemented and fabricated using low-cost printed circuit board (PCB) manufacturing tools.

There is a conflicting requirement between widening the bandwidth of the six-port network while keeping its size small for use as a QPSK modulator. Most of the six-port designs presented in the open literature are not compact in size and many of them have moderate operational bandwidths. For example, in [6] a six-port QPSK modulator operating from 3.1 to 4.8 GHz is presented, but does not have a compact size. The prototype presented in [4, 12] demonstrates operation within a limited 3–4 GHz band. It is also bulky as it uses non-integrated power dividers and couplers that are connected using coaxial cables.

A simulation of a compact wideband six-port modulator was described in [7]; however, it uses a multilayer structure, which poses several fabrication challenges. Misalignment or presence of air gaps between layers can cause performance degradation. To overcome the tight fabrication tolerances faced in a multilayer structure, a six-port network on a single-layer PCB is designed and used here in direct modulation. An example is the six-port network on a single-layer substrate working as a modulator presented in [13]. However, its shortfall is the limited operational bandwidth of 7 to 8 GHz. This paper presents a single microstrip layer six-port network for use in a modulator system that offers an increased operational bandwidth of 66%; from 4.5 to 9 GHz.

2 Principle of operation of a six-port network as a QPSK modulator

The six-port network is a linear network that has six ports; port 1 [local oscillator (LO) port], port 2 (RF out) and ports 3–6 (termination ports). The basic components of the six-port network are power dividers, 3 dB quadrature couplers and phase shifters. Interestingly, the six-port network can be formed in various configurations that are based on different combinations of these components. In general, the six-port network can be divided into four different architecture types that are shown schematically in Fig. 2. These architecture types include the following components:
Type-1: two power dividers, two 3 dB quadrature couplers and one 90° phase shifter (newly proposed here).

Type-2: one power divider and three 3 dB quadrature couplers with one port terminated in a matched load [4–7, 14].

Type-3: four 3 dB quadrature couplers and one 90° phase shifter with two ports terminated in a matched load [15].

Type-4: two power dividers, two 3 dB quadrature couplers and two 45° phase shifters [16].

The operation of the six-port network can be described by its S-parameter matrix which relates the reflected, \( b_i \) and incident waves, \( a_i \) at each port.

Regardless of which configuration of Fig. 2 is chosen, the S-parameter matrix is the same assuming an ideal operation of each component. The relation between the reflected and incident waves in the six-port network is given as [17, 18]

\[
\begin{bmatrix}
    b_1 \\
    b_2 \\
    b_3 \\
    b_4 \\
    b_5 \\
    b_6 
\end{bmatrix} =
\begin{bmatrix}
    \frac{1}{2}(a_3 + ja_4 + ja_5 - a_6) \\
    \frac{1}{2}(a_3 + ja_4 - a_5 + ja_6) \\
    \frac{1}{2}(-a_1 + a_2) \\
    \frac{1}{2}(ja_1 + ja_2) \\
    \frac{1}{2}(ja_1 - a_2) \\
    \frac{1}{2}(-a_1 + ja_2)
\end{bmatrix}
\]

(1)

The final aim is to find a simplified expression for the output signal at port 2, \( b_2 \). Assuming ideal operation at port 2 with no
The four symbols of the modulated QPSK signal shown in Table 1 can be obtained by connecting different combinations of port terminations to ports 3–6. The relative phase differences of each symbol, $\Delta \Phi$, can be obtained by measuring the phase of $S_{21}$ for each symbol, referenced to the phase of $S_{221}$ for symbol 1.

### 3 Six-port network implementation

As described in Section 2, a six-port network can be designed using any of the configurations shown in Fig. 2. The design of type-2, 3 and 4 six-port networks can be found in the open literature [4–7, 14–16] whereas the type-1 six-port network is a novel architecture, newly proposed herein for use as a modulator. The advantage of the type-1 architecture is that it does not require a matched termination of an extra port as in the type-2 and 3 architectures. In comparison with type-4, the proposed device requires only one phase shifter.

In this work, a novel type-1 six-port network is designed, manufactured and tested for use as a QPSK modulator. For comparison, a conventional type-2 six-port network is also designed, manufactured and tested. To validate the comparison, both the type-1 and type-2 six-port networks are constructed using the same types of broadband components, namely double-stage Wilkinson power dividers [19], elliptical disk quadrature couplers [20] and in the case of the type-1 six-port network, a broadband-coupled microstrip-CPW 90° phase shifter [21]. The individual layouts for these components are shown in Fig. 4.

These microwave components are broadband and are designed in microstrip technology using a single double-sided PCB substrate. In the undertaken designs, a Rogers RT6010 substrate with dielectric constant of 10.2, thickness of 0.635 mm and loss tangent of 0.0023 is used for the construction of the six-port networks. The simulation tool CST Microwave Studio is used to optimise the design and verify the performance.

The double-stage Wilkinson divider has been designed according to the method described in [19]. The design of the couplers and the phase shifter follow the strategy explained in [20, 21], followed by adjustments of the final component dimensions in CST. The final dimensions of each component are shown in Table 2.

After the optimisation of each component using CST, they are combined to form the type-1 and 2 six-port networks. Fig. 5 shows the photographs of the fabricated six-port network prototypes.

Both structures are very compact, with dimensions of 50 × 70 mm for the type-1 network, and 50 × 65 mm for the type-2 network. The type-1 structure is 5 mm longer because of the additional length of the phase shifter. The required resistor values for the Wilkinson divider shown in Fig. 4a are 91 and 240 Ω and they are realised using...
standard size chip resistors, 0603 (1.6 × 0.8 mm) and 0805 (2.0 × 1.5 mm), respectively. The proposed type-1 and 2 six-port networks have broadband features, and can operate within an octave bandwidth from 4.5 to 9 GHz. They also have the advantage that they can be designed on a single PCB substrate; they are fully planar and compatible with other microstrip circuits.

The performances of the fabricated six-port networks were measured using a HP8510 VNA. With regard to the S-parameters for the ideal cases, each of the two devices should feature high return losses at ports 1–6, high isolation between port 1 and 2 and 6 dB insertion loss from port 1 to ports 3–6 and from port 2 to ports 3–6. Fig. 6 shows both the simulated and measured results for the reflection coefficient at port 1, the transmission coefficients from port 1 to ports 3–6, and isolation between ports 1 and 2.

As observed from Fig. 6, there is a relatively good agreement between the simulated and measured results. Type-1 and type-2 six-port networks have comparable results in terms of the transmission coefficients and reflection coefficients. The simulated and measured transmission coefficients are 6.5 ± 1.5 dB over the frequency band from 4.5 to 9 GHz. It is found that the simulated and measured results are well matched, and the transmission coefficients are close to the theoretical value of 6 dB over the operational frequency band. There are small amplitude imbalances between the transmission coefficients of the order of 2 dB because of the non-ideal behaviour of the elliptical disk couplers, caused by fabrication errors. The simulated and measured reflection coefficient at port 1 is greater than 10 dB over the 4.5–9 GHz band for both devices.

To minimise the effect of the carrier leakage from the LO signal to the RF output in a QPSK modulator, the isolation between ports 1 and 2 should be maximised [18]. For the type-1 six-port network, the isolation is greater than 17 dB over the frequency band from 4.8 to 9 GHz. For the type-2 six-port network, the isolation is greater than 17 dB over the frequency band from 4 to 9 GHz. The slight difference between the simulated and measured results can be attributed to the inaccurate values of the chip resistors of the Wilkinson dividers, and the non-ideal behaviour of the elliptical couplers.

The reflections at each termination are also another cause for carrier leakage [18]. Thus, the return losses at ports 3–6 should be minimised as well. The simulated and measured return losses at each of the termination ports for both devices are greater than 10 dB over the 4.5–9 GHz band (not shown here).

When considering the phase of the transmission coefficients between ports 1 and 2 and the remaining ports, it is important to verify that it is of an appropriate value and that it stays approximately constant as a function of frequency. This is to ensure that the phases also have a broadband feature since this will determine the phases of the QPSK symbols. The phase differences between ports 3 and 4 and between ports 5 and 6 when ports 1 or 2 are designated as the input ports are shown in Fig. 7. Theoretically, the phase difference should be 90° when referring to port 1 and −90° when referring to port 2 as the input port. Both configurations show a constant

**Table 2** Final dimensions of each of the components that comprise the six-port network

<table>
<thead>
<tr>
<th>Component</th>
<th>Parameter</th>
<th>Value</th>
</tr>
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<tbody>
<tr>
<td>double-stage Wilkinson power divider</td>
<td>W1</td>
<td>0.16 mm</td>
</tr>
<tr>
<td></td>
<td>W2</td>
<td>0.37 mm</td>
</tr>
<tr>
<td></td>
<td>l1</td>
<td>1.5 mm</td>
</tr>
<tr>
<td></td>
<td>l2</td>
<td>1.0 mm</td>
</tr>
<tr>
<td>elliptical disk quadrature coupler</td>
<td>A</td>
<td>5.41 mm</td>
</tr>
<tr>
<td></td>
<td>B</td>
<td>3.41 mm</td>
</tr>
<tr>
<td></td>
<td>Ls</td>
<td>3.85 mm</td>
</tr>
<tr>
<td></td>
<td>Ws</td>
<td>1.88 mm</td>
</tr>
<tr>
<td></td>
<td>Φ</td>
<td>87.4°</td>
</tr>
<tr>
<td>broadside-coupled microstrip-CPW 90° phase shifter</td>
<td>Dm</td>
<td>3.5 mm</td>
</tr>
<tr>
<td></td>
<td>Dc</td>
<td>1.5 mm</td>
</tr>
<tr>
<td></td>
<td>Ds</td>
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</tr>
<tr>
<td></td>
<td>l1</td>
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<td>4.1 mm</td>
</tr>
<tr>
<td></td>
<td>l3</td>
<td>4.9 mm</td>
</tr>
</tbody>
</table>

As observed from Fig. 6, there is a relatively good agreement between the simulated and measured results. Type-1 and type-2 six-port networks have comparable results in terms of the transmission coefficients and reflection coefficients. The simulated and measured transmission coefficients are 6.5 ± 1.5 dB over the frequency band from 4.5 to 9 GHz. It is found that the simulated and measured results are well matched, and the transmission coefficients are close to the theoretical value of 6 dB over the operational frequency band. There are small amplitude imbalances between the transmission coefficients of the order of 2 dB because of the non-ideal behaviour of the elliptical disk couplers, caused by fabrication errors. The simulated and measured reflection coefficient at port 1 is greater than 10 dB over the 4.5–9 GHz band for both devices.

To minimise the effect of the carrier leakage from the LO signal to the RF output in a QPSK modulator, the isolation between ports 1 and 2 should be maximised [18]. For the type-1 six-port network, the isolation is greater than 17 dB over the frequency band from 4.8 to 9 GHz. For the type-2 six-port network, the isolation is greater than 17 dB over the frequency band from 4 to 9 GHz. The slight difference between the simulated and measured results can be attributed to the inaccurate values of the chip resistors of the Wilkinson dividers, and the non-ideal behaviour of the elliptical couplers.

The reflections at each termination are also another cause for carrier leakage [18]. Thus, the return losses at ports 3–6 should be minimised as well. The simulated and measured return losses at each of the termination ports for both devices are greater than 10 dB over the 4.5–9 GHz band (not shown here).

When considering the phase of the transmission coefficients between ports 1 and 2 and the remaining ports, it is important to verify that it is of an appropriate value and that it stays approximately constant as a function of frequency. This is to ensure that the phases also have a broadband feature since this will determine the phases of the QPSK symbols. The phase differences between ports 3 and 4 and between ports 5 and 6 when ports 1 or 2 are designated as the input ports are shown in Fig. 7. Theoretically, the phase difference should be 90° when referring to port 1 and −90° when referring to port 2 as the input port. Both configurations show a constant...
phase shift for both simulated and measured results over the intended frequency band of 4.5–9 GHz. The phase shift is close to the theoretical value with a phase imbalance of $+10^\circ$ within the intended frequency band.

According to the simulated and measured results, the type-1 six-port network has comparable results to the type-2 six-port network. Both of the proposed six-port networks exhibit the required $\pm 90^\circ$ phase characteristics for use in the modulator systems.

4 QPSK modulation results using the proposed six-port networks

The operation of the six-port modulator is simulated using a circuit envelope simulator in agilent’s advanced design system (ADS). The device’s scattering parameters obtained from the electromagnetic simulations in CST microwave studio and from the measurements are used to model its behaviour in ADS. The block diagram of both of the six-port modulators for simulation in ADS is shown in Fig. 8.

As shown in Fig. 8, the six-port QPSK modulator consists of a six-port network, four single pole double throw (SPDT) switches, LO and vector signal generator. The six-port network block can be either type-1 or 2 six-port networks. The input port of the six-port network is connected to the LO. Ports 3, 4, 5 and 6 of the six-port network are connected to four SPDT switches. Each output of the SPDT switches is connected to open and short terminations. The vector signal generator provides I and Q baseband signals that control the SPDT switches. The switches of ports 3 and 4 are controlled by the I signal whereas the switches of ports 5 and 6 are controlled by the Q signal.
To demonstrate the wideband operation of the six-port QPSK modulator, the envelope simulation is performed at three representative LO frequencies; 5, 6.5 and 8 GHz. Fig. 9 shows the constellation diagrams of the QPSK symbol using the simulated and measured $S$-parameters at 5, 6.5 and 8 GHz for both types of six-port networks, using a symbol rate of 400 Msymbols/s, which is a high symbol rate used in new-generation high-capacity communication systems.

As can be observed from the Fig. 9, the constellation diagrams using the simulated and measured responses of the six-port networks are similar to the theory described in Section 2. All the QPSK symbols shown in Fig. 9 should ideally follow the values of $b_2$ shown in Table 1. In terms of the phase, all the symbols are very similar to the theoretical ones which are located at $45^\circ$, $135^\circ$, $225^\circ$ and $315^\circ$. In terms of the amplitude, a notable difference between each symbol can be seen, mainly when using the measured $S$-parameters, because of the imperfect isolation between ports 1 and 2 and increased reflections at ports 3–6. It is to be noted that the phase difference between different symbols and not their amplitude is the main parameter used in the modulation/demodulation of QPSK systems.

To verify the operation of the six-port modulator over the intended frequency band, the phase characteristics of each symbol are plotted against frequency. The modulators are tested using coaxial open-circuit and short-circuit terminations connected to ports 3–6, as shown in Fig. 3. As the six-port networks are designed to operate over a wide frequency band, the open-circuit and short-circuit...
terminations should also have a wideband feature. For this reason, the wideband open-circuit and short-circuit terminations manufactured by Fairview Microwave Inc. are used with the six-port modulator. The open (SC2165-SMA male) and short (SC2133-SMA male) feature constant reflection coefficients in the range of 2–18 GHz. The six-port networks are connected to the open and short terminations following the different combinations shown in Table 1, and their $S_{21}$ characteristics are measured. The relative phase differences of each symbol, $\Delta \Phi$, are obtained by subtracting the phase of $S_{21}$ of each symbol from the phase of $S_{21}$ of symbol 1.

Fig. 10 presents the simulated and measured phase of each QPSK symbol over the 4–9 GHz band for the type-1 and type-2 modulators, respectively. Ideally, the phase difference for each symbol should be $-90^\circ$ ($270^\circ$), $90^\circ$, and $180^\circ$ between symbol 1 and symbols 2, 3, 4, respectively, as shown in Table 1. The simulated phase difference between any pair of neighbouring symbols is $90^\circ \pm 20^\circ$ for the type-1 modulator and $90^\circ \pm 20^\circ$ for the type-2 modulator across the band from 4.5 to 9 GHz. The measured results show a phase difference of $90^\circ \pm 20^\circ$ for the type-1 modulator and $90^\circ \pm 20^\circ$ for the type-2 modulator between any pair of neighbouring symbols across the band from 4.7 to 9 GHz. These phase difference values are sufficient to correctly decode the symbols at the receiver.

5 Conclusion

In this paper, the operation of a compact wideband six-port network operating as a QPSK modulator in the frequency band from 4.5 to 9 GHz has been presented. The modulation has been accomplished using two types of single-layer six-port networks which use broadband power dividers, couplers and phase shifters. The QPSK symbol analysis of the designed modulators has been simulated using ADS and measured using wideband open and short terminations. The presented six-port modulator offers accurate QPSK symbol modulation at a symbol rate of 400 Msymbols/s across an octave bandwidth from 4.5 to 9 GHz (66% operational bandwidth), which is suitable for use in new-generation high-capacity communication systems.

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7 References


Wideband Planar Crossover Using Two-Port and Four-Port Microstrip to Slotline Transitions

A. M. Abbosh

Abstract—The design of a wideband crossover that includes a pair of two-port and another pair of four-port microstrip-slotline transitions is presented. The utilized transitions are designed such that the resultant planar crossover has high isolation and return loss, and low insertion loss and deviation in the group delay across a wideband. The simulated and measured results of a developed 9 mm x 13 mm crossover on a substrate of 10.2 dielectric constant show less than 0.5 dB insertion loss, more than 15 dB return loss, more than 15 dB isolation and less than 0.1 ns deviation in the group delay across the band from 4.8 to 7.2 GHz (40% fractional bandwidth).

Index Terms—Crossover, four port devices, microstrip components.

I. INTRODUCTION

CROSSOVERS are needed in monolithic microwave integrated circuits, microwave multichip modules, antenna distribution networks and many more due to the ever-increasing complexity of modern systems. The traditional approach to realizing crossovers is to use air-bridge bonds. However, these bonds lead to a non-planar structure, and thus, an increased fabrication cost and complexity. To minimize the cost and maintain the planar structure of microwave circuits, four-port devices that allow a pair of intersecting lines to cross each other, while maintaining the required isolation between the two signal paths, are increasingly needed.

Different configurations for crossovers, such as modified ring structures, microstrip to coplanar waveguide transitions or cascaded couplers, have been proposed [1]–[5]. The performance presented in [1] indicates a maximum achievable fractional bandwidth of around 20%. The device proposed in [2] achieves around 40% bandwidth, but with more than 1 dB insertion loss across that band due to the use of extremely narrow lines. The four-section branch-line structure in [3] shows 33% fractional bandwidth based on 20 dB isolation as a reference with up to 1 ns deviation in the group delay, whereas a three-section structure in [4] achieves 5% bandwidth at dual bands. The crossover in [5] is proposed for high power application with 13% fractional bandwidth.

In this letter, a combination of two- and four-port microstrip-slotline transitions is utilized to build a wideband planar crossover. The transitions and the connecting lines are designed for a high isolation between two signal paths with low insertion loss at each of them. The proposed method is validated via simulations and measurements.

II. DESIGN

The proposed crossover is shown in Fig. 1. It is designed to enable the signal entering port #1 to reach port #3 with minimum insertion loss and maximum isolation from the other signal passing in the intersecting path that connects port #2 to port #4. The crossover is comprised of a pair of two-port microstrip-slotline transitions and a pair of four-port microstrip-slotline transitions. As required in planar circuits, the four ports of the crossover are located at the top layer of the substrate. The slotlines needed to realize the required characteristics are located in the bottom layer that includes the ground plane.

The two-port transitions depicted in Fig. 1 are formed by two complementary structures. One of them is a microstrip line terminated with a capacitive circular disk of radius \( r_m \), whereas the other structure is a slotline terminated with an inductive circular slot of radius \( r_s \). The two parts of the transition are electromagnetically coupled at their intersection. The equivalent circuit for the two-port transition can be approximated by the diagram shown in Fig. 2(a), where \( L_s \) and \( \theta_s \) are the inductance of the circular slot stub and its effective electrical length, respectively, whereas \( C_{m} \) and \( \theta_m \) are the capacitance and effective electrical length of the circular microstrip stub, respectively.

With a proper choice of the position and radii of the circular stubs, \( L_s \) and \( C_m \) become negligible. If \( \theta_s \) and \( \theta_m \) are equal...
to quarter of the effective wavelength, the circular slot stub becomes a virtual open circuit, whereas the circular microstrip stub becomes a virtual short circuit. Thus, a simplified model that only includes the transformer’s ratio \( n_1 \) and the characteristic impedances of the microstrip line \( (Z_m) \) and the slotline \( (Z_s1) \) can be used to predict the performance and the initial dimensions. The ratio \( n_1 \) depends on the widths of the microstrip and slot lines at the coupling region and characteristics of the substrate [6]–[8]. In order to have a transition with no insertion loss, \( Z_m \) and \( Z_s1 \) should have the relation \( Z_m = n_1^2 Z_s1 \).

The equivalent circuit for the other main component of the proposed crossover, i.e. the four-port microstrip-slotline transition, is depicted in Fig. 2(b). That transition is designed here to enhance the isolation between the two transmission paths, Port1-Port3 and Port2-Port4. To achieve that target, the transformer’s ratio \( n_2 \) is designed to be small by reducing the coupling area relative to the substrate’s thickness [8]. The width of the microstrip line at the coupling region \( w_3 \) is narrowed. Also, since the transformer’s ratio \( n_2 \) has its maximum value at the normal intersection between the slot and microstrip lines, the intersection is inclined in the utilized four-port transitions by using an elliptical slot oval (Fig. 1).

To further reduce the coupling at the four-port transitions, the slotline at the four-port transitions is designed to have low characteristic impedance \( (Z_s2) \) by using a narrow-width slotline \( (S_1) \). In this case, the transmission coefficient between the very low impedance of the slotline as seen at the microstrip side \( (n_2^2 Z_s2) \) and the microstrip line impedance at the transition is small. To avoid introducing any manufacturing difficulties in the design, the dimensions in the coupling area \( (w_2 \text{ and } S_1) \) are kept practical.

To eliminate the effect of the non-zero coupling at the four-port transitions on the isolation, the distance \( l_1 \) shown in Fig. 1 is chosen to be half of the guide wavelength at the center of the required band. This choice enables the destructive combination of any signals that are coupled to/from the slotline from/to the microstrip line at the two four-port transitions.

In order to explain the operation of the proposed crossover, the equivalent circuit of the whole structure is derived from Figs. 1, 2 and shown in Fig. 3. If a signal enters port 1 (P1), a small fraction of that signal is coupled to point A due to the small, but non-zero coupling of the first four-port transition. The same value of signal is also coupled to point A from the second four-port transition. However, this coupled signal is out-of-phase with the other one due to the additional trip along the 180° microstrip line. Thus, the combination of the two coupled signals at A is zero. Effectively, zero net signals are coupled to the ports P2 & P4, and thus, the whole signal emerges from P3.

If a signal enters P2, that signal is coupled to the slotline due to the perfect matching of the first two-port transition. A small fraction of that signal is then coupled to the microstrip line connecting P1 to P3 at the two four-port transitions. The combination of those coupled signals travelling to P1 or P3 is equal to zero due to the 180° phase shift introduced by the microstrip line \( l_1 \). Thus, the whole signal that enters P2 emerges from P4.

To compensate for the loading effect of the utilized transitions on the return loss of the microstrip and slotlines, the microstrip line \( l_1 \) between the two four-port transitions is linearly tapered. A similar action is taken for the slotline oval extending between the two- and four-port transitions.

III. RESULTS AND DISCUSSION

The proposed crossover is designed to operate across the band from 4 to 8 GHz. Rogers RT6010 with thickness 0.635 mm, dielectric constant 10.2 is used as the substrate. In order to ease the manufacturing process, the minimum value for the width of any slot in the structure is not allowed to be below 0.1 mm, whereas the minimum width of any microstrip line is 0.2 mm.

As a compromise between the requirements of a perfect matching at the two-port transitions, a reasonable isolation at the four-port transitions, and easy to manufacture dimensions, the main design parameters are chosen as \( n_1 = 0.8 \), \( n_2 = 0.5 \), \( Z_{s1} = 80 \Omega \), \( Z_{s2} = 40 \Omega \), and \( Z_m = 50 \Omega \). The widths of the slotlines \( (S_1, S_2, S_3) \), and microstrip lines \( (w_1, w_2, w_3) \) can then be calculated [8], [9]. The radii of the microstrip and slot circles \( (r_m \text{ and } r_s) \) are chosen to be around twice of the microstrip width \( w_2 \) [10]. The calculated values in (mm) of the design parameters are: \( w_1 = w_2 = w_3 = 0.6 \), \( w_4 = 0.3 \), \( l_1 = 9.5 \), \( S_1 = S_2 = 0.1 \), \( S_3 = 0.2 \), and \( r_m = r_s = 1.2 \). The lengths \( l_2 \) and \( l_3 \) are not critical parameters in the design. They are chosen to be very small for a compact structure.

The overall dimensions of the crossover are optimized using CST Microwave Studio. The optimized dimensions (mm) are: \( w_1 = 0.8 \), \( w_2 = 0.5 \), \( w_3 = 0.3 \), \( w_4 = 0.45 \), \( l_1 = 9 \), \( l_2 = 2.1 \), \( l_3 = 9.5 \), \( S_1 = 0.1 \), \( S_2 = 0.13 \), \( S_3 = 0.29 \), and \( r_m = r_s = 1.4 \). The optimized dimensions are generally close to the calculated values.

The overall size of the manufactured crossover (inset of Fig. 4) excluding the input/output feeders is 9 mm × 13 mm. It is worth mentioning that the size can be further reduced, if needed, by replacing the microstrip line \( l_1 \) by a meandered line,
and thus, the size of the slot oval in the ground plane can be reduced consequently.

The simulated and measured performances of the designed crossover are shown in Figs. 4 and 5. The results reveal an insertion loss of less than 0.5 dB across the band from 4.6 to 7.3 GHz (45% fractional bandwidth) in the simulations and from 5.1 to 7.2 GHz in the measurements. The isolation between the two sets of ports and the return loss of the four ports are more than 10 dB, whereas the insertion loss is less than 1.6 dB across the whole investigated band from 4 to 8 GHz. The simulated and measured results agree well with each other.

In order to show the low level of distortion introduced by the developed crossover, the variation of the group delay is depicted in Fig. 4. It is clear that the peak-to-peak variation is less than 0.1 ns across the whole investigated one octave band. This low value in the deviation is compared favourably with the recently designed crossovers that have around 1 ns deviation.

In comparison with the published crossovers, it is possible to claim that the presented design achieves larger fractional bandwidth based on the 0.5 dB insertion loss reference. It also has smaller deviation in the group delay across that band.

IV. CONCLUSION

A wideband planar crossover that uses two- and four-port microstrip-slotline transitions has been presented. The measured results of the developed device show less than 0.5 dB of insertion loss, more than 15 dB of isolation and return loss, and less than 0.1 ns deviation in the group delay across 40% fractional bandwidth.

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fields inside the cavity can be expressed in terms of TM\(_{mn0}\) modes.

The microstrip square patch of Figure 1 with the ground plane can be represented as a dielectric loaded cavity. The top and bottom of that cavity are represented as perfect electric walls. Assuming that the thickness of the substrate is very small compared with the wavelength, the tangential magnetic fields at the edges of the cavity are very small, and the field variation along the normal axis (z-axis) is considered constant. Moreover, the fringing fields along the edges of the patch are also very small, and thus, the electric field is nearly normal to the surface of the patch. Thus, the side walls of the patch can be represented as perfect magnetic walls and the only field in the cavity is the transverse magnetic (TM) field configuration. Following the analysis presented in [4], the electric (\(E\)) and magnetic (\(H\)) fields inside the cavity can be expressed in terms of TM\(_{mn0}\) modes.

\[
E_z = A_1 \cos(m \pi x/l) \cos(n \pi y/l) \quad (1a)
\]
\[
H_x = A_2 \cos(m \pi x/l) \sin(n \pi y/l) \quad (1b)
\]
\[
H_y = A_3 \sin(m \pi x/l) \cos(n \pi y/l) \quad (1c)
\]

\(A_i\): Amplitude of the fields, \(l\): Length of the patch, \(m\) and \(n\): Mode numbers.

From (1), it is possible to show that the two fundamental modes of the structure (TM\(_{100}\) and TM\(_{010}\)) have the following resonant frequency

\[
f_r = \frac{c}{2l \sqrt{\varepsilon_r}} \quad (2)
\]

\(\varepsilon_r\): Dielectric constant of the substrate, \(c\): Speed of light in free-space. The field distributions for the two fundamental modes are found from (1). For the mode TM\(_{100}\)

\[
H_y = A_3 \sin(m \pi x/l); H_x = 0 \quad (3)
\]

For the mode TM\(_{010}\)

\[
H_x = A_2 \sin(n \pi y/l); H_y = 0 \quad (4)
\]

The above results indicate that the two fundamental modes have orthogonal magnetic fields. According to Poynting vector theory, microwave signals flow in the direction defined by the cross-product of the electric and magnetic fields. Thus, the signal flows in the \(x\)-direction for the TM\(_{100}\) mode and in the \(y\)-direction for the TM\(_{010}\) mode. Thus, each face-to-face pair of ports (Port\#1 and 3, or 2 and 4 in Fig. 1) can be properly aligned in the manner depicted in Figure 1 to couple one of those modes. In this case, the isolation is high between the two pairs of ports while maintaining a low insertion loss between the face-to-face ports. The four symmetrical slits of rectangular shapes with dimensions \((d_1 \times d_2)\) are cut from the patch at the positions indicated in Figure 1 to further improve the isolation between the two modes.

### 3. RESULTS AND DISCUSSION

The proposed crossover is designed to operate across the WLAN band from 5.15 to 5.85 GHz. Rogers RT6010 with thickness 0.635 mm, dielectric constant 10.2 is used as the substrate. The resonant frequency of the patch is chosen to be the center of the band, that is, 5.5 GHz. The side length of the used square patch is equal to half of the guide wavelength calculated at \(f_r\) as derived from (2). The four ports are designed to have 50 \(\Omega\) impedance. The overall dimensions of the crossover are optimized using CST Microwave Studio. The optimized dimensions (mm) are: \(l = 8.3\), \(d_1 = 3\), \(d_2 = 0.5\), \(w_1 = 0.69\), and \(w_2 = 0.57\).

The simulated and measured performances of the designed crossover are shown in Figure 2. The results reveal an insertion loss of less than 1 dB across the band from 5.1 to 6 GHz in the simulations and from 5.15 to 5.85 GHz in the measured results. The insertion loss is less than 0.5 dB across the band from 5.2 to 5.8 GHz.

The four ports of the device are well matched with more than 25 dB return loss at the center of the band. The measured return loss is more than 13 dB across the band from 5.15 to 5.85 GHz. The isolation between the two sets of ports (1–3 and 2–4) is more than 12 dB across the whole investigated band.
from 5 to 6 GHz. The simulated and measured results agree well with each other.

One of the main contributions of the proposed design is the extremely low level of signal distortion introduced by the crossover. As depicted in Figure 2, the peak-to-peak variation in the group delay is less than 0.06 ns across the whole investigated band. This result is comparable with those achieved by the recent designed crossovers [1–3]. The extremely low deviation in the group delay is a must for a distortionless performance of the modern high-capacity communication systems.

To clarify the operation of the crossover at the two modes, the distribution of the magnetic fields is calculated using the software tool for the proposed crossover at the center frequency (5.5 GHz) and shown in Figure 3. If the input signal is applied at Port#1, the magnetic field in the $x$-direction ($H_x$) has the dominant value [Fig. 3(a)], whereas the field in the $y$-direction ($H_y$) is almost zero [Fig. 3(b)]. Because the electric field is in the $z$-direction as indicated in (1a), the microwave signal flows in the $y$-direction, and thus, the signal is coupled to Port#3. The mode of operation in this case is TM$_{010}$ as represented by (4). If the signal is applied at Port#2, $H_y$ has the dominant value [Fig. 3(d)], whereas $H_x$ is almost zero [Fig. 3(c)]. The signal flows in the $x$-direction and captured by Port#4. The mode of operation in this case is TM$_{100}$ as represented by (3). The simulated variations of the fields at the two modes as shown in Figure 3 agree well with the derived variations in (3) and (4).

4. CONCLUSION

A truncated microstrip patch has been used in the design of a planar microwave crossover. The operation of the proposed crossover is based on generating two orthogonal modes. Each of those modes is used to couple a pair of face-to-face ports. With the proper positioning of the ports, the two modes, and thus, the two pairs of ports are well isolated. A prototype that is designed to cover the WLAN band from 5.15 to 5.85 GHz has proven the validity of the crossover. The insertion loss is less than 1 dB, the return loss is more than 13 dB, and the isolation is more than 12 dB across the whole band. The main features of the proposed crossover are the very low deviation in the group delay (less than 0.06 ns) and the capability to handle a large microwave power due to the used microstrip patch.

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Abstract—The design of an ultra-wideband crossover for use in printed microwave circuits is presented. It employs a pair of broadside-coupled microstrip-to-coplanar waveguide (CPW) transitions, and a pair of uniplanar microstrip-to-CPW transitions. A lumped-element equivalent circuit is used to explain the operation of the proposed crossover. Its performance is evaluated via full-wave electromagnetic simulations and measurements. The designed device is constructed on a single substrate, and thus, it is fully compatible with microstrip-based microwave circuits. The designed device is constructed on a single substrate, and thus, it is fully compatible with microstrip-based microwave circuits. The crossover is shown to operate across the frequency band from 3.1 to 11 GHz with more than 15 dB of isolation, less than 1 dB of insertion loss, and less than 0.1 ns of deviation in the group delay.

Index Terms—Crossover, four-port devices, microstrip components.

I. INTRODUCTION

Crossovers are usually required in printed microwave circuits whenever an intersection between transmission lines carrying different signals at the same layer is inevitable. This situation is a usual scenario in high density monolithic microwave integrated circuits and microwave multichip modules.

The traditional approach to designing crossovers is to use air-bridge bonds [1], or wired vias [2]. However, the air-bridges lead to non-planar structures, and increased fabrication costs and complexity. The use of wired vias leads to an insertion loss that increases progressively with frequency due to the effect of the parasitic elements of those wired vias. An attractive alternative is to use planar four-port devices that allow a pair of intersecting lines to cross each other, while maintaining the required isolation between the two signal paths.

Several methods were proposed for the design of broadband via-less crossovers using planar configurations, such as modified rings [3], [4], microstrip to coplanar waveguide (CPW) transitions [5], cascaded couplers [6], patch resonator [7], or microstrip to slotline transitions [8]. The fractional bandwidths achieved in the double-rings of [3] and modified ring of [4] are 20% and 26%, respectively. 44% bandwidth is obtained from the crossover that uses microstrip to CPW transitions and filters [5], whereas a crossover based on cascaded couplers has 33% bandwidth [6]. A crossover based on patch resonators for distortionless high power applications achieves 14% fractional bandwidth [7]. In [8], two pairs of microstrip to slotline transitions are used to build a crossover with 40% fractional bandwidth.

An ultra-wideband (UWB) crossover operating across more than 110% fractional bandwidth is presented. It is based on using different types of transitions [9]–[14]. The proposed structure is fully planar, and thus, compatible with the microstrip-based printed circuit technology.

II. DESIGN

The structure of the proposed crossover is shown in Fig. 1. The device can be viewed as an assembly of two crossing lines. The first one connects Port 1 to Port 2, whereas the second one connects Port 3 to Port 4. As can be deduced from the structure, the crossover is constructed using the two sides of a single dielectric substrate. It is fully planar as all the ports are located on the top layer of the substrate. Therefore it is suitable for integration with other microstrip-based devices.

The first crossing line depicted in Fig. 1 is implemented using a pair of microstrip-CPW transition technique which is adapted from [9], [10]. The second crossing line incorporated in the crossover to connect Port 3 to Port 4 is implemented using a pair of uniplanar microstrip to tapered CPW transition. This technique transforms the signal entering, for example, Port 3 from the microstrip mode to the CPW mode for an enhanced isolation from the CPW located at the bottom layer.

To explain the operation of the proposed crossover, its equivalent circuit is shown in Fig. 2(a). To get this circuit, it is assumed that the coupled length \(l_2\) is quarter of the effective wavelength \(\lambda_e\) at the center of the band. The distances \(l_4\) and \(l_6\) equal \(l_2\). The two transformers with turn’s ratio \(r_1\) and \(r_2\) represent the electromagnetic coupling between the microstrip patch and the CPW patch of the two utilized transitions. For the tight coupling achieved in the proposed structure, \(r_1\) and \(r_2\) are close to 1. From [15], achieving a value of around 0.95 for \(r_1\) and \(r_2\) is feasible. For a perfect matching at the four ports of the circuit, the following condition should be realized [16] assuming that the four ports of the device have \(Z_{n1} = \Box\) impedance

\[
Z_{n1} = r_1^2 Z_{c1} = r_2^2 Z_{c2} = \Box\]  

(1)

\(Z_{c1}\) and \(Z_{c2}\): impedances of the CPW at the bottom and top layers, respectively. For \(r_1 = r_2 = 1\), hence, \(Z_{c1} = Z_{c2} = \Box\).

The third transformer with turn’s ratio \(r_3\) represents the undesired coupling between the two circuits. Since the signal in the two circuits travels in a CPW mode at the position of the intersection, the value of \(r_3\) is close to zero ensuring a high isolation between the two circuits.

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are also of the slots in the CPWs of the utilized microstrip patches and parasitic inductances (1). The dimensions of the top layer CPW covered the UWB spectrum from 3.1 to 10.6 GHz with more than 25 dB of isolation. It is to be noted, though, that the equivalent circuit, and thus, the ADS simulations do not include the effect of the losses in the substrate, any radiation losses, and the parasitic elements on the overall performance of the crossover is negligible.

The equivalent circuit in Fig. 2(a) includes parasitic capacitances (C1, C2) of the microstrip patches and parasitic inductances (L1, L2) of the slots in the CPWs of the utilized transitions. For properly designed transitions, the effect of those parasitic elements on the overall performance of the crossover is negligible.

The simulation tool Agilent ADS is used to calculate the S-parameters of the equivalent circuit depicted in Fig. 2(a). The elements of the equivalent circuit used in the calculation are indicated in Fig. 2(a) as per the explained design guidelines. The calculated performance depicted in Fig. 2(b) reveals a crossover covering the UWB spectrum from 3.1 to 10.6 GHz with more than 25 dB of isolation. It is to be noted, though, that the equivalent circuit, and thus, the ADS simulations do not include the effect of the losses in the substrate, any radiation losses, and the effect of the tapered ground of the top layer CPW as shown in Fig. 1.

The final step in the design is to find the required physical dimensions for the structure in Fig. 1. The dimensions of the CPW line (s1 and r1) are chosen for 55 Ω impedance as required by (1). The dimensions of the top layer CPW (s2, r5, r7) are also chosen for 55 Ω impedance using the equations of conductor-backed CPW [17]. The width (w1) of the four microstrip ports is chosen for 50 Ω impedance. The initial values for (w2, r3, r5) are determined by the guidelines described in [9], [10]. The overall dimensions of the crossover are optimized using the simulator HFSS. The final dimensions using the substrate RT6010 (rtε = 9.000, tan δ = 0.002) are given in Fig. 1. A sensitivity study shows that the change in the performance is negligible for up to 10% fabrication tolerance in the design parameters.

III. RESULTS AND DISCUSSION

The proposed crossover is tested via full-wave electromagnetic simulations. A prototype is also fabricated (inset of Fig. 3) and tested. The dimensions of the fabricated crossover excluding the feeding ports used for the testing are 8 mm × 15 mm.

The simulated and measured performance is shown in Figs. 3 and 4. The simulated insertion loss is less than 0.7 dB across the band from 3.1 to 11 GHz. The measured insertion loss is less than 1 dB across the same band except at around 9.5 GHz where the loss is around 1.1 dB at one of the crossing lines. The 0.3 dB average difference between the measured and simulated insertion loss comes from the four Sub-Miniature-A (SMA) connectors needed for the experimental test.

As depicted in Fig. 4, the return loss at the four ports is more than 10 dB across the band from 3.1 to 11 GHz. The isolation between the two crossing lines is greater than 15 dB across the same band. Relatively, good agreement between the simulated and measured results is obtained. From comparing Fig. 4

Fig. 1. Configuration of the proposed crossover. (a) Overall structure (dark colour: top layer), (b) top layer, and (c) bottom layer. Final dimensions in (mm): w1 = 0.64, w2 = 4.2, w3 = 4.1, w4 = 0.86, w5 = 5.8, w6 = 0.23, w7 = 1.5, l1 = 3.6, l2 = 1.7, l3 = 1.8, l4 = 8.7, l5 = 3.3, l6 = 7.9, s1 = 0.26, s2 = 0.49.

Fig. 2. (a) Equivalent circuit, and (b) calculated performance for Z = 50 Ω, Z12 = Zc2 = 55 Ω, C1 = C2 = 0.01 pF, Lc1 = Lc2 = 10 nH, n1 = n2 = 0.95, and n3 = 0.1.
To summarize the benefits of the proposed crossover, its properties compared with the recent published via-less structures are shown in Table I. It is clear that the proposed device has much wider fractional bandwidth than the other crossovers. Apart from the extremely low group delay in the narrowband design presented in [7], the proposed crossover has the lowest group delay compared with available data of other wideband crossovers despite its slightly larger size.

IV. CONCLUSION

A compact via-less UWB crossover on a single substrate has been presented. The crossover is based on using broadside-coupled and direct-coupled microstrip-to-CPW transitions. The device has more than 15 dB of isolation and less than 1 dB of insertion loss across the band from 3.1 to 11 GHz. It is fully compatible with printed microstrip circuits.

REFERENCES

input gate capacitance of $M_{\text{inf}}$ is beneficial for adjusting the locked tuning range of ILFT. Therefore, the locked tuning ranges were measured in Figure 7 at various gate voltage of $M_{\text{inf}}$ and the $V_{\text{gs}}$ was fixed at 0.6 V and the injection signal power was −5 dBm. The measured locking ranges can be further enhanced by increasing the gate bias voltage due to the suppression of gate capacitance.

The first issue is that the naturally running frequency of the oscillator must be very close to the desired harmonic of the reference to achieve injection-locking [9]. In Figure 8 different situations, which occur depending on the injection frequency, are shown. When injection frequency is within the lock range, two oscillator spectrums (master and slave) are combined to form a single peak as shown in Figure 8(b). Two sidebands can be observed after injection locking. The magnitude of these sidebands is varied for injection in different frequency offsets from carrier frequency. If the frequency of the slave oscillator is outside of the injection locking range but very close to that, the spectrum shown in Figure 8(c) occurs and this situation is called “quasi lock” and if the frequency of the injected signal deviates far from the lock range, the spectrum shown in Figure 8(d) can happen which is called “fast beat” [10].

4. CONCLUSION

A 77 GHz ILFT has been designed and implemented in 90 nm standard CMOS technology under a 1.2 V supply voltage. The use of slow wave EC-CPW inductor and transmission line-based ILFT shrinks the harmonic generator chip size. In addition, the inverter-core Colpitts differential VCO fundamental signal also demonstrates a wide tuning range. The frequency range of the output signal can be controlled from 76.4 to 79.15 GHz with a tuning range of 2.78 GHz. Due to the low-power and wide-locking-range operation, this approach can be applied to 77 GHz millimeter-wave wireless communication for automotive applications are feasible.

A comparison table of this work and published CMOS injection-locked frequency multipliers (ILFM) at millimeter wave operation frequencies was shown in Table 1. The performance of the proposed 77-GHz ILFT can achieve the lowest power dissipation among the CMOS ILFM circuits.

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WIDEBAND PLANAR MICROSTRIP CROSSEVER WITH HIGH POWER HANDLING CAPABILITY AND LOW DISTORTION

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ABSTRACT: The design of a planar four-port microstrip crossover is presented. The design starts by using a half-wavelength square patch and two sets of orthogonal feeding lines. To achieve a wideband performance, four circular slots are introduced in the square patch. A parametric analysis is used to investigate the effect of the utilized slots. Full-wave simulations and measurements are used to validate the proposed design. The simulated and measured data show less than 1 dB insertion loss, more than 10 dB return loss, more than 19 dB isolation, and about 0.2 ns group delay deviation across the band from 2.1 to 2.75 GHz. © 2012 Wiley Periodicals, Inc. Microwave Opt Technol Lett 55:439–443, 2012; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.27335

Key words: wideband crossover; microstrip crossover; four-port devices

1. INTRODUCTION

Microstrip lines crossover is a common problem in most of the modern microwave integrated circuits. The poor isolation and the high return losses caused by the crossover dramatically reduce the whole circuit performance. An ideal four-port crossover provides 0 dB insertion loss to diagonal ports and perfect isolation to adjacent ports. The most common way to isolate the signals on the intersection is to use a three-dimensional structure, for example, bond wires, air bridges, or multilayer structures [1–3]. The use of those nonplanar structures increases the fabrication complexity and cost significantly.

In the literature, a few fully planar structures were proposed, such as cascaded hybrid and symmetric four-port junctions [4–6]. The achievable bandwidth in the recently proposed crossovers is 5% at dual-bands [5] and 33% by using four cascaded sections. However, the designs that are based on several multi-section structures suffer from a large deviation in the group delay. For example, the deviation in the group delay is around 1 ns in the four-section structure proposed in Ref. 6. That large variation in the group delay prohibits the use of those devices in wideband applications.

We propose in this work a new planar four-port crossover in the form of a square microstrip patch with four arc-shaped slots. The crossover is fed using four microstrip feeders located at the four corners. The direction of each feeding line is perpendicular to the two adjacent feeding lines. The simulated and measured results show that the developed crossover operates across 28% fractional bandwidth with more than 19 dB isolation and less
than 1 dB insertion loss. The deviation in the group delay across that band is around 0.2 ns.

2. PROPOSED PATCH CROSSOVER

The proposed crossover configuration is shown in Figure 1. The starting point in the design is a square microstrip patch with four feeding microstrip lines at the top layer of the substrate. The four feeding lines are arranged such that each one of them is perpendicular to the two adjacent lines. The bottom layer of the substrate includes the ground plane. To enable wideband performance, four arc-shaped slots of radius $R_0$, width $w_1$, and separation distance $w_2$ are embedded in the patch as shown in Figure 1.

To explain the principle of operation of the proposed structure, the cavity theory [7] is used. The structure shown in Figure 1 can be represented as a dielectric loaded cavity under the following conditions. The width of the slots and the thickness of the substrate are very small compared with the wavelength, and the fringing fields along the edges of the patch and those leaked from the slots are also very small. In this case, the field variation along the normal axis ($z$-axis) is almost constant, and the electric field is nearly normal to the surface of the patch. Thus, the side walls of the patch can be represented as perfect magnetic walls and the only field in the cavity is the transverse magnetic (TM) field configuration. Following the analysis presented in Ref. 7, the two fundamental modes of the fields in the structure are TM100 and TM010 with the following resonant frequency

$$f_r = \frac{c}{\sqrt{\varepsilon_r} \lambda}$$  \hspace{1cm} (1)

$L$: Length of the patch, $\varepsilon_r$: Dielectric constant of the substrate, $c$: Speed of light in free-space.

The field distributions for the two fundamental modes are found from (1). For the mode TM100

$$H_z = A_1 \sin(\pi x / l); H_x = 0$$  \hspace{1cm} (2)

For the mode TM010

$$H_x = A_2 \sin(\pi y / l); H_y = 0$$  \hspace{1cm} (3)

$A_i$: amplitude of the field.

The above results indicate that the two fundamental modes have orthogonal magnetic fields. The signal flows in the $x$-direction for the TM100 mode and in the $y$-direction for the TM010 mode. Thus, each face-to-face pair of ports of the structure in Figure 1 can be properly aligned to couple one of those modes. In this case, the structure can be designed to operate as a four-port crossover.

Based on (1), the side length of the patch is equal to half a wavelength. The crossover is designed here to work at the resonant frequency 2.4 GHz using the substrate Rogers RT6010LM (thickness 0.635 mm and $\varepsilon_r$ = 10.2). Assuming 50 impedance for the feeding lines, the dimensions are: $l = 19.8$ mm and $W = 0.6$ mm. The utilized substrate has a side length of $W = 27$ mm.

After finding the values of the main design parameters, the effect of the other three parameters ($w_1$, $w_2$, and $R_0$) is investigated separately to show their impact on the bandwidth.

Figure 2 shows the changes of the return loss and insertion loss as a function of the radius of the circular slots $R_0$ for $w_1$ and $w_2$ equal to 0.5 mm. It is clear from the figure that for large values of $R_0$, the response shows one transmission pole at a position defined by the half wavelength dimensions of the crossover as predicted using the cavity theory. Reducing the value of $R_0$ splits the central pole into two. The distance between those two poles increases with reducing $R_0$. Thus, it is clear that this parameter can be used to increase the bandwidth. There is a limit on the possible increase in the bandwidth as the insertion loss starts to deteriorate for small values for $R_0$ as shown in Figure 2b for $R_0 = 8.2$ mm.

Figure 3 shows the variation in the return loss and insertion loss as a function of the slot width $w_1$ for $R_0 = 8.2$ mm, and $w_2 = 0.5$ mm. The graph shows that decreasing the slot width has a minor effect on the lower resonant frequency by shifting it slightly to higher values. However, $w_1$ has a significant impact on the second resonant frequency. Decreasing $w_1$ shifts that frequency to higher values, which eventually leads to a welcome increase in the fractional bandwidth. It is to be noted from Figure 3 that there is a limit on the increase in the bandwidth that can be achieved as very small value for $w_1$ means deterioration in the performance at the center of the band besides the expected manufacturing problems associated with using very
narrow slots. Thus, it is clear from Figure 6 that the optimum slot width $w_1$ is between 0.4 and 0.6 mm.

Figure 4 shows the changes in the return loss and insertion loss as a function of the separation distance between the two arc slots $w_2$. In this case, increasing the distance between the slots shifts the second resonance frequency to higher values, which leads to an increase in the fractional bandwidth. As for $w_1$, the design parameter $w_2$ has a small effect on the lower resonant frequency. It is clear from Figure 4 that the optimum separation distance $w_2$ is between 0.4 and 0.6 mm.

After using the parametric analysis to estimate the values of the design parameters ($w_1$, $w_2$, and $R_0$) for a wideband performance, the overall structure of the device is optimized for the maximum possible bandwidth that is centered at 2.4 GHz. The optimized values are: $R_0 = 8.2$ mm, $w_1 = 0.5$ mm, and $w_2 = 0.5$ mm.

Having found the optimum dimensions of the crossover, the next step is to present a visualization of the field distribution in the crossover structure as a proof of the utilized cavity theory. The result using the software HFSS is shown in Figure 5 when a 2.4 GHz signal is applied at port #1. It is clear that no signal is coupled to the neighboring ports #2 and 4, whereas the whole input signal is coupled to port #3. The designed crossover is fabricated as shown in Figure 6 on a $27 \times 27$ mm² substrate and tested.

Figure 7 shows the simulated and measured results for the return loss, insertion loss, and isolation of the developed crossover. It is to be noted that due to symmetry $S_{11}=S_{22}=S_{33}=S_{44}$, $S_{13}=S_{24}$, and $S_{12}=S_{34}$. The crossover has less than 0.5 dB insertion loss and more than 19 dB isolation across 28% fractional bandwidth extending from 2.1 to 2.75 GHz. The results show a very good agreement between the simulated and measured results.

For a distortionless performance across a wideband, the group delay of the device should have a flat response across the operational band. To inspect that parameter, the group delay is calculated and measured as depicted in Figure 8. The deviation in the value of the group delay across the 28% band is only 0.2 ns, which is an attractive small value.

It is worth mentioning that the proposed crossover does not have any slots in its ground plane. This feature enables its use and integration in high-density microwave circuits. Moreover, the structure of the crossover, which is in the form of a microstrip patch, can handle the high power levels used in microwave transmitter circuits compared with other designs that utilize narrow transmission lines in their structure. Those properties in addition to the extremely low deviation in the group delay

![Figure 2](image)

**Figure 2** Variation of (a) return loss and (b) insertion loss with $R_0$, for $w_1 = w_2 = 0.5$ mm. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

![Figure 3](image)

**Figure 3** Variation of (a) return loss and (b) insertion loss as a function of $w_1$. ($w_1 = 0.5$ mm, $R_0 = 8.2$ mm). [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
should make the proposed crossover an attractive candidate for wideband microwave circuit applications.

3. CONCLUSION

The design of a planar microstrip crossover has been presented. The proposed crossover utilizes a square microstrip patch with four perpendicular microstrip feeding lines. To enable a wideband performance, four arc-shaped slots are introduced into the square patch. The simulated and measured results have shown less than 1 dB insertion loss, more than 19 dB isolation, and less than 0.2 ns deviation in the group delay across the band from 2.1 to 2.75 GHz.

Figure 4 Variation of (a) return loss and (b) insertion loss as a function of w2. (w1 = 0.5 mm, R0 = 8.2 mm). [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 5 Electric field distribution on the square patch crossover. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 6 Fabricated prototype. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 7 Simulated and measured S parameters of the proposed crossover. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 8 Simulated and measured group delay. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
A NARROW-BAND HIGH-TEMPERATURE SUPERCONDUCTING FILTER WITH WIDE UPPER STOPBAND

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ABSTRACT: This article presents a novel interdigital capacitor resonator (ICR), which has a high spurious frequency and compact size. A six-pole high-temperature superconducting (HTS) bandpass filter (BPF) has been designed and fabricated with this ICR on a doublesided MgO-based YBCO thin film. Measurement at cryogenic temperature shows a fractional bandwidth of 0.4%, and an ultra high spurious passband width is 0.4%. A novel ICR is developed in the design procedure, which has the advantages of high spurious frequency, compact size, and weak coupling.

Keywords: high-temperature superconducting filter; interdigital capacitor resonator; wide stopband; narrowband

1. INTRODUCTION

High-temperature superconducting (HTS) films have a lower surface resistance (about two orders lower at 10 GHz) at cryogenic temperature than conventional metal at room temperature. The unloaded quality factor of a HTS microstrip resonator below 10 GHz is usually above 10000 [1]. Thus, HTS filters have many advantages, such as very low insertion loss, steep band edges, high out-of-band rejection, and so on. HTS filters have received much attention in recent years and have been applied to improve microwave systems, such as mobile communication, radar detection, astronomical observation, and so on [2–5].

Compact size and excellent passband performance are more concerned in the HTS filter design. However, HTS microstrip filters always have unwanted spurious passbands, leading to a degradation of the out-of-band performance. As a matter of fact, wide stopband response is required in many filter applications. One natural way of suppressing the spurious passband is cascading a lowpass filter, but additional insertion loss maybe introduced, and dimension will be enlarged. Quarter wavelength resonator has the nearest spurious frequency at 3f0 [4], but the earthing of resonators on HTS film could affect the passband performance. Recently, stepped impedance resonators (SIRs) and interdigital capacitor resonators (ICRs) are developed to realize wide stopband performance [5–9]. Because SIRs usually have a large line width of the low impedance section to increase the spurious frequency, they are often used in filter design at upper UHF band or even higher frequency to obtain compact size. However, the first spurious frequency of ICRs usually locates at about 3f0, and the fundamental frequency can be reduced by increasing self capacitance, so ICRs are very suitable to develop a microstrip filter with wide stopband performance and compact size at lower frequency.

In this article, theory analysis of ICR is studied; the design and experimental performance of a six-pole HTS BPF at UHF band is presented. The first spurious passband of the filter is located at 2034 MHz (about 4f0/3), of which the fractional bandwidth is 0.4%. A novel ICR is developed in the design procedure, which has the advantages of high spurious frequency, compact size, and weak coupling.

2. RESONATOR DESIGN

An ICR, as shown in Figure 1(a), has many fingers in parallel at the two ends of the resonator, which introduces the loading capacitance. So, a ICR be simply modeled as capacitively loaded transmission line resonator [10, 11] of Figure 1(b), where C1/2 is the loaded capacitance at either end, Z, and d are the characteristic impedance, the propagation constant, and the length of the unloaded line, respectively. The transmission matrix can be written as:

\[
\begin{bmatrix}
V_1 \\
I_1
\end{bmatrix} = \begin{bmatrix}
AB \\
CD
\end{bmatrix} \begin{bmatrix}
V_2 \\
-I_2
\end{bmatrix}
\]

(1)

\[
A = D = \cos \theta - \frac{1}{2} \alpha C_1 Z \sin \theta
\]

\[
B = j Z \sin \theta
\]

\[
C = j \alpha C_L \cos \theta + \frac{1}{2} \sin \theta - \frac{1}{4} \alpha C_1^2 Z \sin \theta
\]

(2)

Figure 1 (a) A novel interdigital capacitor resonator applied in the filter design and (b) capacitively loaded transmission line resonator

A Compact UWB Three-Way Power Divider

Amin M. Abbosh

Abstract—A three-way power divider with ultra wideband behavior is presented. It has a compact size with an overall dimension of 20 mm × 30 mm. The proposed divider utilizes broadside coupling via multilayer microstrip/slot transitions of elliptical shape. The simulated and measured results show that the proposed device has 4.77 ± 1 dB insertion loss, better than 17 dB return loss, and better than 15 dB isolation across the frequency band 3.1 to 10.6 GHz.

Index Terms—Power divider (PD), ultra wideband (UWB), Wilkinson divider.

I. INTRODUCTION

POWER dividers (PDs) are fundamental components of many microwave circuits and subsystems. They are widely used in antenna arrays, power amplifiers, mixers, phase shifters and vector modulators [1].

The simplest three-way PD is the Wilkinson divider [2]. Although it provides a match at all ports and high isolation between output ports, a three-way Wilkinson divider presents serious packaging problems. It requires a 3-D floating common node to connect all isolation resistors together. This requirement makes fabrication difficult and complex, especially in high frequency bands using monolithic microwave integrated circuits [1]. To overcome these problems, a modified circuit, which uses additional isolation resistors and wire bonding, was proposed in [3]. However, the isolation performance of the proposed circuit is very sensitive to the length of the bond wire, and its operation is limited to the lower frequency band.

A new configuration of the three-way divider was introduced in a recent paper [4]. It consists of two resistors and three four microstrip coupled lines. The proposed configuration can modify a three-way Wilkinson PD from a 3-D configuration into a two-dimensional one. However, the measured performance of the divider shows that it has a narrow bandwidth. Furthermore, it requires a narrow spacing between the microstrip lines. This makes the fabrication process difficult, knowing that its performance is sensitive to the coupled lines spacing.

Another type of broadband planar three-way PD is the tapered-line PD [5]. It provides a broadband high-pass characteristic because of the tapered-line impedance transformers. The tapered-line PDs utilize resistive films or strip resistors which cover all or part of the area between the tapered-line conductors to obtain good output isolation. However, those resistors cause a significant insertion loss in the high frequency range. Thus, they degrade the performance of the divider and limit its useful bandwidth [6].

In this letter, the configuration of a compact multilayer three-way PD with ultra wideband (UWB) performance is presented. Simple rules are proposed to design the device. The simulated and measured results show that the insertion loss is equal to 4.77 ± 1 dB for each of the three output ports across the band 3.1–10.6 GHz. The proposed divider exhibits better than 17 dB return loss at its ports with more than 15 dB isolation across the ultra wide frequency band.

II. DESIGN

The configuration of the proposed multilayer three-way PD is shown in Fig. 1. It consists of five conductor layers interleaved by three dielectrics. The input and one of the output ports, which are stripline ports, are located at the mid layer of the structure, while the other two output ports, which are microstrip ports, are at the top and bottom layers. The ground plane, which also includes the coupling slot, is at the second and fourth layers of the circuit. The microstrip coupled patches and the slots are of elliptical shapes, similar to those used to fabricate the UWB directional couplers in [7]. The two isolated ports, which are Ports 5 and 6 in Fig. 1(f), have no power output. They are terminated in matched loads to absorb any reflected signal from the output ports which may degrade the isolation performance of the device.

The mid layer of the proposed divider is considered to be connected to the input port (Port 1) at one side, and one of the output ports (Port 2) at the other side. As the power is required to divide equally between the three output ports, then the coupling between the mid layer and any of the two output ports (Port 3 and Port 4 in Fig. 1) is equal to: \( \epsilon = \sqrt{1/\epsilon_0} = 0.5773 \) (or –4.77 dB). The output power from Port 2 is: \( \sqrt{1 - \epsilon^2} = \sqrt{1/\epsilon_0^2} = 0.5773 \), i.e. insertion loss = 4.77 dB.

The even \( (Z_{oe}) \) and odd \( (Z_{od}) \) mode characteristic impedances for each of the coupled patches are calculated using the following equations:

\[
Z_{oe} = Z_0 \sqrt{\frac{1 + \epsilon}{1 - \epsilon}}; \quad Z_{od} = Z_0 \sqrt{\frac{1 - \epsilon}{1 + \epsilon}} \tag{1}
\]

where \( Z_0 \) is the characteristic impedance of the input/output ports of the coupler. Assuming that \( Z_0 = 50 \) \( \Omega \) and \( \epsilon = \sqrt{1/\epsilon_0^2} \), then: \( Z_{oe} = 96.5 \) \( \Omega \); and \( Z_{od} = 25.9 \) \( \Omega \).

Dimensions of the elliptical microstrips and slots offering the required even and odd mode characteristic impedances can be determined by extending the quasi-static approach presented in [7], [8] to the case of multilayer coupler. That approach was found to give accurate results when the distance between the coupled lines is less than \( \lambda/20 \) [9], where \( \lambda \) is the wavelength inside the used substrate. In this letter, I used a substrate that meets this requirement across the UWB.
Using the quasi-static method, $Z_{\text{in}}$ and $Z_{\text{out}}$ can be proven to be

$$Z_{\text{in}} = \frac{\varepsilon_r K'(k_1)}{\sqrt{\varepsilon_r}} K(k_1), \quad Z_{\text{out}} = \frac{\varepsilon_r K'(k_2)}{\sqrt{\varepsilon_r}} K(k_2)$$

(2)

where $\varepsilon_r$ is the dielectric constant of the substrate, $K(k)$ is the first kind elliptical integral and $K'(k) = K(\sqrt{1-k^2})$.

Rearranging the equations in [7] and [8], it is possible to find the design parameters as (3) and (4), shown at the bottom of the page, where $P_{\text{in}}$ and $P_{\text{out}}$ are the major diameters of the elliptical slot and coupled microstrip, respectively, $P_{\text{sec}}$ is the secondary diameter of the slot and coupled microstrip, $t$ is thickness of the substrate, and $l$ is equal to quarter of the effective wavelength calculated at the centre frequency, which is 6.85 GHz. The relation between $P_{\text{sec}}$ and $l$ is given by [7]

$$P_{\text{sec}} = \left(\sqrt{l^2 + \left(\frac{4.77 + \varepsilon_r \varepsilon_0}{\varepsilon_r} P_{\text{in}} P_{\text{sec}} / l^2\right)} + l\right) / 2$$

(5)

The last step of the design is to calculate the width of the input and output stripline/microstrip ports. They are determined to give 50 $\Omega$ characteristic impedance using the well known stripline/microstrip equations [1].

### III. RESULTS AND DISCUSSION

The validity of the presented design method was tested by building a three-way PD aimed at the operation in the UWB range 3.1 to 10.6 GHz. Rogers RO4003C (with $\varepsilon_r = 3.38$, $t = 0.508$ mm, and loss tangent $\tan \delta = 0.0027$) was selected for the divider’s development. Using the proposed design method and with the help of the optimization capability of the software Ansoft HFSSv10, parameters of the coupler were found to be: $P_{\text{in}}$ for the top and bottom layers $= 4.8$ mm, $P_{\text{out}}$ for the mid layer $= 4.4$ mm, $P_{\text{sec}} = 9$ mm, $P_{\text{sec}} = 6.8$ mm, width of the stripline input/output port $= 0.64$ mm, and width of the microstrip output ports $= 1.2$ mm. It was found that the optimized values of the design parameters are less than 10% different from those obtained by the described design method. This indicates the high accuracy of the method. A photograph of the developed device is shown in Fig. 2. It has a compact size with an overall dimension of 20 mm $\times$ 30 mm.

The designed PD was tested via simulation and measurements. The simulation was performed using the commercial software Ansoft HFSSv10, whereas the measurements were done using a vector network analyzer.

The insertion loss of the three output ports, as shown in Fig. 3, is equal to $4.77 \pm 1$ dB (ideal value $4.77$ dB) across the band 3.1 to 10.6 GHz revealing an UWB performance. The return loss for the input/output ports of the device is shown in Fig. 4 (note that because of symmetry, $S_{22} = S_{11}$ and $S_{44} = S_{33}$). It is better than 17 dB for all the ports of the device. The isolation between the three output ports is presented in Fig. 5 (due to symmetry $S_{32} = S_{42}$). The simulated and measured isolation between ports 3 and 2 (or between the ports 4 and 2) is better than 19 dB, whereas it is better than 17 dB between the

\[
\begin{align*}
\frac{1}{\varepsilon_r} & = \frac{\sinh^2\left(\frac{\pi t}{2.4} \frac{P_{\text{in}} P_{\text{sec}}}{l}\right)}{\frac{\pi t}{2.4} \sinh^2\left(\frac{\pi t}{2.4} \frac{P_{\text{in}} P_{\text{sec}}}{l}\right)} + \frac{\cosh^2\left(\frac{\pi t}{2.4} \frac{P_{\text{in}} P_{\text{sec}}}{l}\right)}{\frac{\pi t}{2.4} \cosh^2\left(\frac{\pi t}{2.4} \frac{P_{\text{in}} P_{\text{sec}}}{l}\right)}
\end{align*}
\]

(3)

\[
\begin{align*}
\frac{1}{\varepsilon_r} & = \frac{\sinh^2\left(\frac{\pi t}{2.4} \frac{P_{\text{in}} P_{\text{sec}}}{l}\right)}{\frac{\pi t}{2.4} \sinh^2\left(\frac{\pi t}{2.4} \frac{P_{\text{in}} P_{\text{sec}}}{l}\right)} + \frac{\cosh^2\left(\frac{\pi t}{2.4} \frac{P_{\text{in}} P_{\text{sec}}}{l}\right)}{\frac{\pi t}{2.4} \cosh^2\left(\frac{\pi t}{2.4} \frac{P_{\text{in}} P_{\text{sec}}}{l}\right)}
\end{align*}
\]

(4)
ports 3 and 4 (or between the ports 5 and 6) across the band 3.1–10.6 GHz.

The simulated and measured results in Fig. 5 also show that there is negligible coupling between the top and bottom layer. The two ports of the top layer (ports 4 and 5) are well isolated from the two ports of the bottom layer (ports 3 and 6) by more than 15 dB across the whole band.

Concerning the phase performance of the splitter, the measured and simulated results (not shown here) indicated that the output signals from ports 3 and 4 are in phase and they are different by 90° ± 0.5° from that of port 2.

IV. CONCLUSION

A three-way PD with UWB behavior has been presented. It has a compact size with an overall dimension of 20 mm × 30 mm. The proposed divider utilizes broadside coupling via multilayer microstrip/slot transitions of elliptical shape. The simulated and measured results of the developed device have shown equal three-way power division, good return loss, and isolation over the band 3.1–0.6 GHz.

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accurate FDTD methods owing to the asymmetry of their time-stepping schemes are depicted. It can be found that they are also same.

4. CONCLUSION
In this article, one of staggered multistep schemes, that is, the staggered Adams-Bashforth scheme, is used to construct an explicit fourth-order accurate FDTD method. The comparison of the FDTD method using the SAB scheme with the FDTD method using the SBD scheme shows that the performance of their numerical dispersions is same but the stability restraint of the former is relaxed by 33.3%. In addition, the FDTD methods using staggered multistep schemes are not difficult to implement. If their application is combined with some new technology such as memory-efficient formulation technology, they will become more absorbing.

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AN UWB PLANAR OUT-OF-PHASE POWER DIVIDER EMPLOYING PARALLEL STRIPLINE-MICROSTRIP TRANSITIONS

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ABSTRACT: The design of a planar out-of-phase (180°) power divider with an ultra wideband performance is presented. The device employs two dielectric substrates with a common ground plane. A transition from a parallel stripline to two microstrip lines is formed to divide power equally and with 180° phase difference from a stripline input port to two microstrip line output ports. The simulated and measured results of the proposed divider show equal power division with high stability of phase, good return loss at the three ports, low insertion losses, and fine isolation between the two output ports across the band from 5 to more than 11 GHz. © 2007 Wiley Periodicals, Inc. Microwave Opt Technol Lett 49: 912–914, 2007; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.22324

Key words: microstrip line; stripline; ultra wideband; power divider

1. INTRODUCTION
Power dividers are fundamental components extensively used in a variety of microwave circuits such as amplifiers, mixers, modulators, phase shifters, and antenna array feed networks [1]. According to the phase difference between two output ports, the power divider can be classified as an in-phase or out-of-phase. In the latter case, a 180° phase difference between the output ports is implemented. One example of such an out-of-phase power divider for use in a push–pull type transistor amplifier was described in Ref. 2. In this case, the Wilkinson out-of-phase divider was employed. In general, the Wilkinson divider that is commonly used in various power dividing/combing circuits is classified as in-phase or out-of-phase [2–4]. Because of the use of resistive elements, it offers good impedance match at its three ports accompanied by high isolation between the two output ports. However, the resistive elements also contribute to some insertion loss, which can be up to 1 dB in an operational band. To achieve a wideband performance, a few sections, each of about one quarter wavelength (at the centre frequency of the intended band), have to be employed. This makes this device of a relatively large size in comparison with the operational wavelength.

In this paper an alternative configuration of a planar out-of-phase divider, accompanied by simple design rules, is presented. The device is compact in size and does not use any resistive elements. From the inherent properties of a lossless three-port [1] it does not offer the same quality of return loss at its three ports and isolation between output ports as the Wilkinson divider. However, as shown via computer simulations and experiment, its performance is very reasonable and may be found sufficient in many applications. The return loss at the input port is in the order of 15 dB across an ultra wide frequency band from 3 to more than 11 GHz. The return loss at its output ports and the isolation between them are in the order of 10 dB across the same band.

2. DESIGN
The configuration of the proposed power divider is shown in Figure 1. The device uses two substrates supported by a common ground plane. Its input port (also named E port) is formed by a parallel strip line. Next, a transition from a parallel strip line to two microstrip lines is formed. This arrangement enables an even signal division in magnitude but with 180° phase difference between two microstrip lines. Note that in the parallel strip line region, the common ground plane is removed. However, it exists in the region of the two microstrip lines. Details of this transition can be better viewed in Figure 1(b). To form two output ports, which are suitably separated, the two microstrip lines are bent. To eliminate a path for a DC signal, two capacitors are used. For microwave frequencies, 1 nF chip capacitors offer this function with a minimum insertion loss. In the configuration shown in Figure 1, 1 nF chip capacitors are placed in the two microstrip lines just before the connection of the two output ports. The use of these DC blocks is necessary as one of the strips has to act as a ground. The two microstrip lines (as well as the remaining microstrip lines in the divider structure of Fig. 1) are assumed to have 50 Ω characteristic impedance. Therefore, for a given substrate their width w can be determined using the standard formula [1], which is rewritten here in Eqs. (1) and (2).

\[
Z_o = \frac{60}{\sqrt{\epsilon_{\text{r}}} \ln \frac{8h}{w_o}} \quad \text{for } w_o/h \leq 1, \quad (1)
\]

\[
Z_o = \frac{120\pi}{\sqrt{\epsilon_{\text{r}}} \left[ w_o/h + 1.39 + 0.67 \ln(w_o/h + 1.44) \right]} \quad (2)
\]
for \( w_s/h \gg 1 \), where \( h \) is thickness of the substrate and \( \varepsilon_{me} \) is the effective dielectric constant for the transmission line and it is given by:

\[
\varepsilon_{me} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2\sqrt{1 + 12h/w_m}}
\]

(3)

where \( \varepsilon_r \) is the dielectric constant of the substrate.

Width \( w_s \) of the parallel strip line is also chosen to give 50 \( \Omega \) impedance at the input port, i.e. the \( E \)-port. This can be done using another formula from [1], which is rewritten as Eq. (4).

\[
w_s = \frac{2.4\pi h}{\sqrt{\varepsilon_r}}.
\]

(4)

When calculating width of the strip line using Eq. (4), the thickness \( h \) is equal to twice of the substrate thickness. This is because the two parallel strip lines are on top and bottom of the structure use identical substrates.

As the width required for the parallel strip line \( (w_s) \) is usually wider than that for the microstrip line \( (w_m) \) an elliptical arc is proposed to be used to make the smooth transition. It can be found from computer simulations that the length of the transition of the order of a quarter wavelength (at the centre frequency) can provide good return loss at the input (\( E \)) port.

3. RESULTS

The above outlined design method was applied to a 180° power divider, which would cover the ultra wideband (UWB) frequency range from 3.1 to 10.6 GHz. Rogers RT6010 with thickness 0.64 mm, dielectric constant 10.2, and tangent loss 0.0023 was assumed as a substrate. Photo for the developed device is shown in Figure 2. The overall dimension of the structure is 2 cm \( \times \) 2 cm.

The operation of the device was simulated using the finite element analysis. Next, the device was developed and tested using a vector network analyzer. Note that in this case, SMA coaxial connectors were included in the three ports of the divider. Results of the performed simulations and measurements are shown in Figures 3 and 4. Those results reveal that the power is equally divided between the two output ports with an insertion loss less than 0.3 dB across the band 3 to more than 11 GHz. The return loss at the input port is better than 15 dB for the whole band. At the same time, return loss at the output ports is in the range of 10 dB.
for the same band. Because the device is virtually lossless, isolation between the output ports is sacrificed and is in the range of 10 dB. The measured and simulated difference in phase between the two output ports (not plotted here) is $180^\circ \pm 0.5^\circ$ over the same band. The results presented in Figures 3 and 4 reveal quite a good agreement between the simulated and measured results.

4. CONCLUSIONS

The design of a planar UWB out-of-phase ($180^\circ$) power divider that involves two substrates supported by a common ground plane has been presented. The device employs a transition from a parallel strip line to two microstrip lines. The simulated and measured results of this device show a low insertion loss and good return losses and isolation over the band 3 GHz to more than 11 GHz.

One has to note that in the presented design, the input port ($E$) is of the parallel strip line type. However, it can easily be converted to the microstrip type. This can be accomplished by gradually increasing the width of one of the strips to form a fin shaped ground plane, and by adjusting the width of the second strip to obtain the required characteristic impedance of 50 $\Omega$.

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Abstract—The design of a compact out-of-phase uniplanar power divider operating over an ultra wide frequency band is presented. To achieve an out-of-phase signal division over a large frequency range, a T-junction formed by a slotline and a microstrip line accompanied by wideband microstrip to slotline transitions is employed. The simulated and experimental results of the developed divider show a low insertion loss and good return loss performance of the three ports across the band 3.1–10.6 GHz.

Index Terms—Power divider, ultra wideband (UWB), Wilkinson divider.

I. INTRODUCTION

P

ower dividers are fundamental components extensively used in a variety of microwave circuits such as balanced mixers, modulators, phase shifters, and antenna array feed networks [1]. The simplest type of power divider is a T-junction. It is a three-port network with one input port and two output ports. According to the phase between output ports, this type of power divider can be an in-phase or out-of-phase. In the latter case, a 180° phase difference between the output ports is offered. This phase difference is required in some applications such as a push–pull type amplifier, where the two transistors are fed 180° out-of-phase.

In [2], the authors have shown that the out-of-phase type Wilkinson divider can be designed using parallel strip transmission lines (PSTLs). The symmetrical characteristic of PSTL implies that the “ground” and “signal” lines can be freely interchanged. This leads to the possibility of having in-phase or out-of-phase signal division. One extra step required to use this arrangement is a PSTL to microstrip transition, which the authors of [2] also included in their design. By using this approach, a three-stage Wilkinson divider operating from 1 to 8 GHz was demonstrated. The shortcoming of the presented configuration is that it is not suitable for implementations in which the microwave device requires a uniplanar microstrip arrangement. This arrangement is required for creating an efficient heat sink.

In this letter, the configuration of a compact uniplanar out-of-phase divider with a low insertion loss is presented accompanied by simple design rules. Opposite to the Wilkinson divider, this device does not use any resistive elements. Because of the inherent properties of a lossless three-port [1], which are governed by unitary properties of its scattering matrix, it cannot offer a perfect match at its three ports, as its counterpart with resistors. In addition, isolation between its two output ports is compromised by the quality of match of its input and output ports. The better the match at the input and output ports, the worse is the isolation between the output ports.

In the presented design, the three-port exhibits return losses at its ports in the order of 10 dB across an ultra wide frequency band from 3.1 to 10.6 GHz, as demonstrated via simulations and measurements. The isolation between the output ports is about 8 dB across the same band. This return loss and isolation performance may be found sufficient in many applications even those involving a push–pull type of amplifier.

II. DESIGN

The configuration of the proposed power divider is shown in Fig. 1. The three ports of the divider are at the top layer of the printed circuit board (PCB) while the ground plane is at the bottom layer of the circuit. There is a slot in the shape of a narrow rectangle ended with two circles in the middle of the ground plane. This slot is responsible for guiding the wave from the input port to the output ports.

Prior to its design, it is important to understand the operation of this device. The divider utilizes the series type T-junction formed by a slotline and two arms of a microstrip line. The inherent property of this junction is that the signals coupled from the slotline to the two arms of the microstrip line are of equal magnitude but their phases differ by 180°. In order to efficiently (without reflections) couple a signal from the slotline to the microstrip, the end of the slotline needs to be compensated with an inductive element. Here, it is chosen in the form of a circular slot. In order to convert the input port from the slot type to the microstrip type, a wideband slot to microstrip transition needs...
to be employed [3]. As seen in Fig. 1, the chosen transition is formed by two complementary structures. One includes a microstrip line terminated with a capacitive circular disk and the other one is a slotline terminated with an inductive circular slot. The two are electromagnetically coupled.

Having established the principles of operation of the power divider, a simple procedure can be applied to its design. The width \( \frac{\lambda}{4} \) of the input and output microstrip ports is determined assuming 50 \( \Omega \) characteristic impedance. The distance \( \frac{\lambda}{4} \) between the input and output microstrip line does not have to be fine tuned. Here, it is chosen to be a quarter of the effective wavelength at the center frequency of operation \( f_c \). As a result, the length of the slot \( l = \frac{\lambda}{4} + 2r \).

In order to achieve a high return loss at the microstrip ports, the slot width \( s \) should be chosen to give impedance close to 50 \( \Omega \) as seen from the microstrip side. At the same time, the slot width should not be too narrow to avoid problems with the manufacturing errors. To meet this requirement, a slot with acceptable width and suitable impedance transformation ratio [3] can be used. The equations in [3] and [4] can be employed to accomplish this task. In the present design, the slot width is chosen to give impedance 90 \( \Omega \) and the impedance transformation ratio is 0.8. As the result, the impedance of the slot as seen from the microstrip line is \( (90 \times 0.8^2 = 55 \Omega \).\)

Radius \( r \) of the microstrip and slot circles terminating the microstrip and slot lines are chosen to be around twice of the microstrip width \( \frac{\lambda}{4} \). This choice conforms to the guidelines for designing wideband microstrip/slotline transitions [3].

### III. RESULTS AND DISCUSSION

The above outlined design method was applied to design an out of phase (180°) power divider that would cover the ultra wideband (UWB) range from 3.1 to 10.6 GHz. Rogers RO4003C with thickness 0.508 mm, dielectric constant 3.38 and tangent loss 0.0023 was used as a substrate. Values of the design parameters \((\frac{\lambda}{4}, d, r, \text{ and } s)\) obtained using the outlined design procedure (without any involvement of sophisticated full EM analysis and design packages) were 1.2 mm, 7.3 mm, 2 mm, and 0.2 mm, respectively. According to these dimensions the proposed divider is compact and its length is around quarter of the effective wavelength in microstrip. This compares favourable against the Wilkinson divider presented in [1] where the length was larger than three quarters of the wavelength.

Having determined all of the parameters of the divider, the next step concerned the testing of its performance. The simulated results of the proposed power divider obtained using Ansoft HFSSv9.2 are shown in Fig. 2. These results reveal that the power is equally divided between the two output ports with an insertion loss less than 0.5 dB across the band 3.1 to 10.6 GHz. Also the return loss for the input port and two output ports (note that because of symmetry, \( S_{33} = S_{22} \) and thus \( S_{33} \) is not shown explicitly) is in the range of 10 dB for the whole band. Because of good quality return loss at the three ports, isolation between the output ports is sacrificed and is in the range of 8 dB. This is the result of the earlier discussed unitary properties of the scattering matrix of a lossless three port [1]. The difference in phase between the two output ports (not plotted here) is 180°±0.25° over the same band.

The power divider was developed and tested experimentally. Photograph of the developed power divider is presented in Fig. 3. Measurement results are shown in Fig. 4. The presented results confirm the UWB behaviour of the designed divider. Also they show good agreement with the simulated results. Small discrepancies can be explained by the use of coaxial ports in the experiment. As observed in Fig. 4, the power is equally divided be-
between the two output ports with average insertion loss less than 0.5 dB. This insertion loss is lower than that of the Wilkinson divider presented in [1] where 0.8-dB insertion loss was noted. The return loss for the input port and output ports is in the order of 10 dB and isolation between the two output ports is between 7.5 and 11.5 dB across the band. Note that the presented measured results include losses of coaxial connectors that were used in the experimental testing.

The measured difference in phase between the two output ports is 180°±0.5° over the band 3–11 GHz, as shown in Fig. 5. This proves that the proposed divider is an out-of-phase type.

IV. CONCLUSION

A simple method has been presented to design a compact UWB out-of-phase power divider. The proposed divider is of the uniplanar type and uses complementary structures in the form of microstrips and slots to achieve its UWB performance. The measured results of the developed power divider have shown a low insertion loss, less than 0.5 dB, a good return loss and isolation with high phase stability over the band 3.1–10.6 GHz.

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DESIGN OF A UWB PLANAR 180° HYBRID EXPLOITING MICROSTRIP-SLOT TRANSITIONS

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ABSTRACT: The article describes the design of a planar 180° hybrid with an ultra wideband (UWB) performance. The device employs two substrates with a common ground plane and various microstrip-slot transitions to achieve in-phase and out-of-phase signal division over an ultra wide frequency range. At the initial stage, simple design guidelines are used but the final dimensions are determined using a full-wave analysis and design software package. The simulated and measured results of the proposed device reveal a well balanced power split accompanied by a very good approximation of ideal 180° and 0° differential phase shift across the band 3.1 GHz to more than 11 GHz. Also, low insertion losses, good return loss at all of the four ports, high isolation between the input ports and fine isolation between output ports are noted across this frequency band. © 2007 Wiley Periodicals, Inc. Microwave Opt Technol Lett 49: 1343–1346, 2007; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.22472

Key words: 180° hybrid; microstrip line; stripline; ultra wideband

1. INTRODUCTION

The 180° hybrid, also called magic-T, is a four-port network that offers out of phase (180° phase shift) and in-phase (0° phase shift) signal/power division between its two output ports [1]. The port enabling the out-of-phase signal division is called the Δ-port (delta...
The alternative names for the $\Delta$ and $\Sigma$ ports are E and H ports, respectively. The property of the ideal 180° hybrid is that all four ports are perfectly matched and the $\Sigma$ and $\Delta$ ports, as well as the two output ports, are isolated. These properties of the 180° hybrid make it very attractive for applications in many microwave communication and measurement sub-systems. It is used in antenna feeds, amplifier combiners, balanced mixers, and modulators.

Many of these applications require the 180° hybrid to be integrated with other microwave circuitry. In this case, a planar (microstrip or stripline) realization of this device is essential. One of the well-known planar realizations of the 180° hybrid is a 1.5 Λ ring coupler also called a rat-race hybrid. This device offers about 20–25% bandwidth and hence its application is limited to narrow-band applications. One technique to overcome this shortcoming is to replace its 3Λ/4 transmission line section by a 1/Λ coupled-line [2]. Known as the 180° reverse-phase hybrid-ring coupler, the new device is less frequency sensitive and offers an operational bandwidth in the order of 40% (or one octave). More recent work described in Ref. 3 indicates that by incorporating crossovers, the ring hybrid developed in coplanar waveguide technology not only can deliver more than one octave bandwidth but also can be miniaturized to 0.67 guided wavelengths. However, one has to note that the use of crossovers make the new device nonplanar. The solutions offered in Refs. 2 and 3 provide a 180° hybrid with an increased operational bandwidth. However, none of them enables operation over an ultra wide frequency band. In 2002 US-FCC released an UW frequency band from 3.1 to 10.6 GHz for various applications [4]. The release of this frequency band has sparked a renewed interest in various broadband passive and active components, and antennas. It is because UW technology shows promise to offer new capabilities with respect to such applications as wireless communications [5] and microwave imaging [6].

Out of many configurations of 180° hybrid/magic-T described in the open literature, the ones worthwhile of revisiting with respect to UW applications are those described by de Ronde in [7], and Aikawa and Ogawa in [8]. The magic-T proposed by de Ronde in [7] is formed by a combination of planar in-phase and out of phase microstrip/slotline dividers. In particular, utilization of a slot to microstrip transition enables an UW out-of-phase (180°) signal division due to an inherent property of this series type junction. The in-phase signal division realized by the shunt microstrip junction also offers an UW operation. However, the assembly of the two junctions, resulting in the magic-T, offers only an octave band performance.

The solution presented in Ref. 8 by Aikawa and Ogawa seems to be more advantageous in terms of preserving UW performance of two individual in-phase and out-of-phase power dividers when combined into magic-T. The authors of Ref. 8 presented three complementary configurations, which can offer a UW operation. The proposed solution utilizes slot-to-microstrip or slot-to-microstrip transitions accompanied by either coupled microstrip lines or coupled slotlines. By using a transition between a single slotline and a coupled microstrip line (or alternatively between a microstrip line and coupled slot lines), a fine quality $\Delta$ ($E$) port is realized. In turn, the $\Sigma$ ($H$) port uses a slot to microstrip transition (or a microstrip to slotline transition). Because of the use of coupled lines, a much more amiable assembly of the in and out-of-phase power dividers forming the magic-T junction is realized in comparison with the one obtained in Ref. 8. Our design also features other differences. It incorporates a 3 dB slot-microstrip coupler and a transition between a parallel strip line and two microstrips to achieve good performance of the magic-T in terms of return loss of various ports and isolation between output ports across UW.

2. DESIGN

The configuration of the proposed UWB 180° hybrid is illustrated in Figure 1. The device uses two substrates supported by a common ground plane and various microstrip-slot transitions. It’s $H/\Sigma$ port is formed by a slot-to-microstrip transition. In turn, the $E/\Delta$ port is created by a parallel strip to two microstrip lines transition. When the slot line forming $H/\Sigma$ port is removed, the modified device represents a 180° power divider. In this divider, the connection between a parallel stripline and two microstrip lines serves the purpose of dividing the signal launched at port $E$ equally in manufacturing tolerances. The coupled lines are edge-coupled lines and in order to offer a 180° phase shift they have to be in a very close proximity. For the manufactured device, the measured best insertion loss occurred at the centre frequency of 6 GHz and was 0.7 dB, and the coupling imbalance was 0.9 dB at the same frequency.

In this article, an alternative configuration of a planar 180° hybrid, is presented. In contrast to [8] the out-of-phase division is accomplished using broadside, instead of edge coupled microstrip lines. Because of this choice, a better quality 180° phase shift is expected in comparison with the one obtained in Ref. 8. Our design also features other differences. It incorporates a 3 dB slot-microstrip coupler and a transition between a parallel strip line and two microstrips to achieve good performance of the magic-T in terms of return loss of various ports and isolation between output ports across UWB.
magnitude but out-of-phase between the two microstrip lines. Note that in the parallel strip line region, the common ground plane is removed. However, it exists in the two microstrip lines region. Details of this transition are better viewed in Figure 1(b). Next, the two microstrip lines are connected to two elliptical conducting discs. The two discs are coupled by a slot, which is formed in the common ground plane. The use of this coupling structure serves the purpose of obtaining better quality return loss and isolation between the two output ports. Note that the device can also function without including this coupling structure.

It is assumed that the two microstrip lines (as well as the remaining microstrip lines in the structure of Fig. 1) are of 50 Ω characteristic impedance. Therefore, for a given substrate their width \( w_m \) is determined using the standard formula [1]. In turn, width \( w_s \) of the parallel strip line is chosen to give 50 Ω impedance using another formula from [1]. When calculating width of the strip line its thickness is assumed to be twice the substrate thickness. As the width required for the parallel strip line (\( w_s \)) is usually wider than that for the microstrip line (\( w_m \)) then an elliptical arc is used to make the smooth transition. The length of this transition needs to be about half wavelength (at the centre frequency) to provide good return loss of the input (E/Δ) port.

The microstrip-slot coupler formed by the two elliptically shaped conducting discs and the slot forming are designed to obtain a 3 dB coupling [9, 10]. This task is accomplished by properly choosing dimensions of the elliptical discs and slot (\( l, D_1, D_2, s, \) and \( r \)). The method to obtain these dimensions has been described in Refs. 9 and 10. The design is based on simple design guidelines followed by a full-wave analysis offered by commercially available software package.

Having designed the out-of phase divider, the inclusion of the in-phase divider to form magic-T is straightforward. As seen in Figure 1, a slotline to microstrip transition accomplishes this task. Using this transition, power delivered to the \( H/\Sigma \) port is equally and in-phase divided between the two parallel microstrip lines. This arrangement offers a better quality phase balance then the configuration with two edge-coupled lines described in Ref. 8. Width of the slot (\( s \)) is chosen so that the input impedance of the \( H/\Sigma \) port is 50 Ω. In this case the formulae described in Refs. 11 and 12 are used. The remaining task is to provide a high return loss at the \( H/\Sigma \) port. This is accomplished by employing a circular slot with radius \( r \) terminating the slotline.

3. RESULTS

The above outlined method was applied to design and build a 180° hybrid divider operating over the UWB frequency range from 3.1 to 10.6 GHz. Rogers RO4003C with thickness 0.508 mm, dielectric constant 3.38, and tangent loss 0.0023 was used as a substrate. The design process was aided with a full electromagnetic simulation package while the measurements were accomplished using a vector network analyzer. Values of the design parameters shown in Figure 1 (\( w_m, w_s, l, D_1, D_2, s, \) and \( r \)) are 1.2, 4.1, 7.4, 4.8, 7.2, 0.1, and 2.5 mm, respectively. The overall dimension of the structure is 3 \( \times \) 4.5 cm

Results of the performed simulations and measurements when the E-port is excited are shown in Figure 2. Those results reveal that when E/Δ port is excited the power is equally split between the two output ports with an insertion loss less than 0.5 dB across the band 3.1 to more than 11 GHz. Also the return loss for the E/Δ port (\( S_{11E} \)) is better than 13 dB for the whole band. The difference in phase between the two output ports (not plotted here) is 180° ± 0.25° over the same band. Good agreement can be noticed between the measured and simulated results.

Figure 3 shows the results when the \( H/\Sigma \) port is excited. In this case, the power is equally divided between the two output ports with an insertion loss less than 0.6 dB across the band 3.7 to more than 11 GHz. The return loss of the \( H/\Sigma \) port (\( S_{11H} \)) is better than 10 dB for the whole band. The difference in phase between the two output ports (not plotted here) is ±0.25° over the same band.

The isolation between the two input (\( E \) and \( H \)) ports was also measured and was found to be better than 25 dB for the whole band. The isolation between the two output ports was found to be better than 8 dB for the same band.

4. CONCLUSION

The design of a UWB 180° hybrid in the planar format involving two substrates supported by a common ground plane has been presented. In order to achieve a UWB performance, the device employs a microstrip-slot coupler and transitions from a parallel strip line to two microstrip lines and from a slotline to two
Gain equalization methods in erbium-doped fiber amplifiers

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ABSTRACT: We show experimental performance comparison of different equalization methods in erbium-doped optical amplifiers based on silica and ZBLAN host materials. Different configurations are analyzed, including the analysis of hybrid silica/ZBLAN configurations, exploring gain, bandwidth, and noise figure performance. © 2007 Wiley Periodicals, Inc. Microwave Opt Technol Lett 49: 1346–1349, 2007; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.22445

Key words: erbium-doped fiber (EDF); optical fiber amplifiers; optical fiber communication; wavelength-division-multiplexing (WDM)

1. INTRODUCTION

Gain equalization in optical amplifiers is a critical issue in the case of wavelength-division-multiplexing optical communication systems. This is because it is necessary to provide a flat gain [1] over the whole amplification bandwidth.

Fiber optic amplifiers (erbium-doped fiber amplifiers and Raman amplifiers) have been extensively studied as key devices for wavelength-division-multiplexing (WDM) optical communication systems with the development of high-power semiconductor laser diodes. The distributed EDFA’s of relatively high optical signal-to-noise ratio (OSNR) and low-noise figure amplification characteristics compared with lumped-type EDFA’s have been demonstrated recently by Kawakami et al., who have obtained good performances of bit-error rate with a hybrid configuration consisting of distributed and lumped EDFA’s [2]. Especially distributed-type Raman amplifiers have attracted considerable attention these days as an important amplifier with EDFA showing a low-noise figure and high OSNR in long-haul transmission systems [3], even though they require high-power optical pumps. By using multiwavelength Raman lasers, it is possible to provide equalization over the amplification spectrum [4]. However, the high optical pumps involve safety problems and damage components as isolators.

In this work, we are going to focus on a comparative study on the performance of different equalization methods in erbium-doped optical amplifiers based on silica and ZBLAN fibers and configurations composed of both of them. A comparison between the topologies has been carried out and the major drawbacks and strong points considering the gain, bandwidth, and the noise figures obtained have been reported.

2. GAIN-EQUALIZATION METHODS OF ERBIUM-DOPED FIBER AMPLIFIERS

One of the major problems in EDFA’s is their nonflat gain profile. The differences in gain among the maximum and the minimum peaks can be as high as 5 dB. This is a problem in wavelength-division-multiplexing systems, compared with configurations designed to amplify only one wavelength. Thus, it is important to provide a flat gain band in the amplification range.
An UWB Planar Out-of-Phase Power Divider Employing Microstrip-Slot and Parallel Stripline-Microstrip Transitions

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Abstract—The design of a planar out-of-phase power divider in microstrip/parallel stripline technology for ultra wideband (UWB) applications is presented. The device employs two substrates with a common ground plane. A coupling between two microstrip lines on the top and bottom substrate layers through a slot in the ground plane, and a suitable transition from two microstrip lines to a parallel stripline are used to achieve an UWB operation. The simulated and measured results of the proposed divider show low insertion losses, good return loss at the three ports, fine isolation and the required 180º phase difference for signals emerging from the two output ports across the band 3.1 to 10.6GHz.

Index Terms—Microstrip line, stripline, ultra wideband, power divider.

I. INTRODUCTION

A power divider formed by a three-port network with an input port and two output ports is a fundamental component extensively used in a variety of microwave circuits [1]. According to the phase difference between the output ports, this device can be classified as a quadrature (90º), in-phase (0º) or out-of-phase (180º).

While most of the early dividers used the waveguide structure, the increasing use of planar technology has led to the development of a wide range of planar dividers such as the Wilkinson divider, the branch line hybrid and the coupled line divider. Wilkinson divider is the most commonly used device to achieve in-phase or out-of-phase signal division [2] - [5]. Because of the use of resistive elements, this type of divider offers good impedance match at its three ports and high isolation between the two output ports. However, the resistive elements also contribute to some insertion loss, which in some applications is unwelcome. In order to achieve a wideband performance a few sections, each of about one quarter wavelength (at the centre frequency) has to be used. This makes the Wilkinson divider of a relatively large size in comparison with the operational wavelength.

In this paper an alternative configuration of a planar out-of-phase divider, accompanied by simple design rules, is presented. The device does not use any resistive elements and achieves an ultra wideband performance using a very compact size. From the inherent properties of a lossless three-port [1] it does not offer the same quality of return loss at its three ports and isolation between output ports, as the Wilkinson divider. However, as shown via computer simulations and measurements its return loss and isolation performance is very reasonable. The return loss at the input port is in the order of 15dB across an ultra wide frequency band from 3.1 to 10.6GHz. The return loss at its output ports is in the order of 10dB while the output ports isolation is about 12dB on the average across the same band. This may be found sufficient in many useful applications.
II. DESIGN

The configuration of the proposed power divider is illustrated in Fig.1. The device uses two substrates supported by a common ground plane. The first section of the power divider includes an input port (E port), which is formed by a parallel strip line. Next, a transition from a parallel strip line to two microstrip lines is formed to divide the signal equally in magnitude but out-of-phase. Note that the common ground plane exists in the two microstrip lines region but is removed in the parallel strip line section. Details can be viewed in Fig.1 b. Next, the two microstrip lines are bent so they can be connected to two elliptically shaped conducting discs, which together with a slot in the common ground plane form a coupling structure.

![Diagram of power divider](image)

Note that the device can also function without the disc-coupling structure. However, its use improves isolation between the two output ports and their return loss.

In order to eliminate a path for a DC signal, two capacitors are used in the output ports. For microwave frequencies, 1nF chip capacitors offer this function with a minimum insertion loss. In the configuration shown in Fig.1, 1nF chip capacitors are placed in the two microstrip lines just before the connection of the two output ports. The use of these DC blocks is necessary as one of the strips in the parallel stripline section has to function as ground.

The two microstrip lines (as well as the remaining microstrip lines in the divider structure of Fig.1) are assumed to have 50Ω characteristic impedance. Therefore, for a given substrate their width $w_m$ is determined using the standard formula [1], which is rewritten here in (1), (2).

$$Z_0 = \frac{60}{\sqrt{\varepsilon_{me}}} \ln(\frac{8h}{w_m} + \frac{w_m}{4h})$$  \hspace{1cm} (1)

for $w_m / h \leq 1$ and;

$$Z_{o_o} = \frac{120\pi}{\sqrt{\varepsilon_{me}}[w_m / h + 1.39 + 0.67ln(w_m / h + 1.44)]}$$  \hspace{1cm} (2)

for $w_m / h \geq 1$.

where $h$ is thickness of the substrate and $\varepsilon_{me}$ is the effective dielectric constant for the transmission line and it is given by;

$$\varepsilon_{me} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2\sqrt{1 + 12h / w_m}}$$  \hspace{1cm} (3)

where $\varepsilon_r$ is the dielectric constant of the substrate.

Width $w_s$ of the parallel strip line is chosen to give 50Ω impedance using another formula from [1], which is rewritten as (4).

$$w_s = \frac{2.4\pi h}{\sqrt{\varepsilon_r}}$$  \hspace{1cm} (4)
When calculating width of the strip line using (4) the thickness $h$ is equal to twice of the substrate thickness. This is because the two parallel strip lines are on top and bottom of the structure formed by two identical substrates.

As the width required for the parallel strip line ($w_p$) is usually wider than that for the microstrip line ($w_m$) then an elliptical arc is used to make the smooth transition. It can be found from computer simulations that the length of the transition of the order of a quarter wavelength (at the centre frequency) can provide good return loss of the input ($E$) port.

The second part of the divider is formed by a microstrip-slot coupler. In this coupler, the microstrip lines are connected to two elliptically shaped conducting discs/patches. The coupling between those two patches is controlled using an elliptical slot in the common ground plane. Dimensions of the three ellipses shown in Fig.1 are chosen to obtain a 3dB coupling between the top and bottom layers. To this purpose, the method described in [4] can be applied. This method initially determines the dimensions of the rectangular shaped slot and the discs. In the next step, the equivalence, in terms of the occupied area, is used to determine the size of ellipses. The initial values for the design parameters ($l$, $D_1$ and $D_2$) shown in Fig.1a are calculated using (5) to (9).

$$k_2 = \tanh(\pi D_2 / 4h) \tag{8}$$

$$l = \frac{c}{4f \sqrt{\varepsilon_{me}}} \tag{9}$$

The final dimensions of the two elliptical patches and the slot can be determined using an iterative process involving the finite element analysis. The aim of this process is to obtain a 3dB coupler and a transition from a parallel strip line to two microstrip lines with high quality performance over an UWB frequency range. Using this approach, usually only a few iterations are required to obtain a satisfactory performance of the divider.

### III. RESULTS

The above outlined design method was applied to a 180º power divider, which would cover the UWB frequency range from 3.1 to 10.6GHz. Rogers RT6010 with thickness 0.64mm, dielectric constant 10.2 and tangent loss 0.0023 was assumed as substrate. Values of the design parameters shown in Fig.1a ($w_m$, $w_s$, $l$, $D_1$ and $D_2$) were 0.6mm, 2mm, 3.7mm, 2.4mm and 4.2mm respectively. The overall size of the structure is 2.5cm×3cm. A photograph of the developed device is shown in Fig.2.

![Fig.2 Photo for the developed device.](image)

The design was assisted with a finite element analysis method while the measurements were performed using a vector network analyser. Results of the simulations and measurements are shown in Fig.3. These results reveal that the...
power is equally divided between the two output ports with an insertion loss less than 0.5dB across the band 3.1 to more than 10.6GHz. Also the return loss for the input port is better than 15dB for the whole band. At the same time, return loss of the output ports is in the range of 10dB for the same band. Because the device is virtually lossless, isolation between the output ports is sacrificed and is in the range of 12dB. The difference in phase between the two output ports (not plotted here) is\(180° \pm 0.5°\) over the same band.

IV. CONCLUSION

The design of an UWB out-of-phase power divider in the planar format involving two substrates supported by a common ground plane has been presented. In order to achieve an UWB performance, the device employs a microstrip-slot coupling arrangement and a transition from a parallel strip line to two microstrip lines. The simulated and measured results of this device show a low insertion loss and good return losses and isolation over the band 3.1 to 10.6GHz.

In the presented design, the input port \(E\) is of the parallel strip line type. However, it can easily be modified to the microstrip type. This can be accomplished by gradually increasing the width of one of the strips to form a fin shaped ground plane and by adjusting the width of the second strip to obtain the required characteristic impedance of 50\(\Omega\).

REFERENCES


Fig.3 Simulated and measured performance of the proposed out-of-phase power divider: (a) coupling and isolation (b) return loss of the ports.
Design of Ultra-Wideband Three-Way Arbitrary Power Dividers

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Abstract—A method to design arbitrary three-way power dividers with ultra-wideband performance is presented. The proposed devices utilize a broadside-coupled structure, which has three coupled layers. The method assumes general asymmetric coupled layers. The design approach exploits the three fundamental modes of propagation: even–even, odd–odd, and odd–even, and the conformal mapping technique to find the coupling factors between the different layers. The method is used to design 1:1:1, 2:1:1, and 4:2:1 three-way power dividers. The designed devices feature a multilayer broadside-coupled microstrip-slot-microstrip configuration using elliptical-shaped structures. The developed power dividers have a compact size with an overall dimension of 20 mm × 30 mm. The simulated and measured results of the manufactured devices show an insertion loss equal to the nominated value ±1 dB. The return loss for the input/output ports of the devices is better than 17, 18, and 13 dB, whereas the isolation between the output ports is better than 17, 14, and 15 dB for the 1:1:1, 2:1:1, and 4:2:1 dividers, respectively, across the 3.1–10.6-GHz band.

Index Terms—Broadside coupling, power divider, three-way power divider, ultra-wideband (UWB).

I. INTRODUCTION

THREE-WAY power dividers are an important part of many microwave systems such as antenna feeders and power amplifiers. The simplest and most popular three-way power divider is the Wilkinson divider [2]. It is a circularly symmetric power divider, which splits an input signal into equal output signals with a good match at all the ports and a high isolation between the output ports. However, the three-way Wilkinson divider presents serious packaging problems. It requires a 3-D floating common node to connect all isolation resistors together. This requirement makes the fabrication difficult and complex, especially at the high-frequency bands [1]. Moreover, the three-way Wilkinson divider is a narrowband device. Hence, it is not suitable for ultra-wideband (UWB) applications, where a good performance is required across the 3.1–10.6-GHz band. To overcome the fabrication difficulties of the Wilkinson three-way divider and to increase the bandwidth, several configurations were proposed [3]–[8]. However, the results shown in those papers indicate a narrowband performance with around 30% fractional bandwidth, and they are all for equal power division.

In [9], a method to design an \( \eta \)-way power divider with an arbitrary power split ratio was introduced. According to the proposed method, the divider consists of \( \eta \) transmission lines with a quarter-wave-long and \( \lambda \) resistors. The device should be terminated in arbitrary impedances to improve the matching and isolation of the different ports. This creates a serious realization problem: to achieve the desired performance, too high or low impedance of the transmission lines needs to be used. Moreover, the measured result of the designed divider shows only 30% fractional bandwidth, which makes them unacceptable for the UWB applications.

The design presented in [10] is a branch-line multistage multiway power divider. It consists of two or more output lines branching in parallel from an input line. The idea was originally proposed in [11]. Multisection transformers were used to achieve a broadband matching between different sections of the device. The design presented in [10] and the other similar designs shown in [12] and [13] suffer from many serious manufacturing and performance problems: very low impedances are required to achieve a good matching between the different sections of the divider, and a poor isolation was noticed between the output ports. Moreover, the developed devices exhibit an acceptable power distribution only across less than 65% fractional bandwidth. The branch-line three-way divider presented in [14] uses realizable values for matching impedances, but it has less than 50% fractional bandwidth.

The broadband high-pass characteristic of the tapered-line impedance transformers was utilized in [15] to build a three-way power divider. It uses resistive films or strip resistors, which cover all or part of the area between the tapered-line conductors, to obtain a good output isolation. However, those resistors cause a significant insertion loss at the high-frequency range. Thus, they degrade the performance of the divider and limit its useful bandwidth.

A modified configuration of the Wilkinson three-way power divider was introduced in [16]. The proposed configuration can transform the three-way Wilkinson power divider from a 3-D configuration into a 2-D one. However, the measured result of the divider reveals a narrowband performance. Furthermore, it requires a narrow spacing between the coupled microstrip lines. This makes the fabrication process difficult, knowing that its performance is sensitive to the coupled lines spacing.

Recently in [17], a multilayer, broadside-coupled, and elliptical-shaped microstrip-slot-microstrip configuration was used to build UWB equal-power three-way divider. The configuration used in [17] was originally proposed by the author to de-
design UWB directional couplers [18]. The three-way divider presented in [17] is composed of three broadside-coupled layers, i.e., the mid layer, which contains the input port and the direct connected output port, the top layer, which contains one of the coupled output ports, and the bottom layer, which contains the other coupled output port. The coupling between the mid and top layers was assumed to be equal to that between the mid and bottom layers. The top and bottom layers were assumed to be perfectly isolated. With these assumptions, it was possible to approximate the three coupled layers with a separate pair of two coupled layers. A simple method based on the two coupled-lines theory was used to calculate the required odd and even impedance for the different coupled layers. Due to this, the method is limited to the special case of an equal power output from the coupled ports.

In this paper, a method to design arbitrary three-way power dividers with an UWB performance is presented. The design method exploits the three fundamental modes of propagation in three coupled lines (even–even, odd–odd, and odd–even) [19]–[25] and the conformal mapping technique [26] to find the coupling factors between the three coupled layers. The method considers the general case of asymmetrical coupled lines. The isolation between the coupled output ports can be calculated using the proposed method. Three devices with different power ratios were designed and manufactured. The simulated and measured results show that the insertion loss is equal to the nominated value ±1 dB for each of the three output ports across the 3.1–10.6-GHz band. The proposed dividers exhibit better than 17-dB return loss at their input/output ports with more than 15-dB isolation between their output ports across the ultra-wide frequency band.

II. ANALYSIS OF THE BROADSIDE-COUPLED STRUCTURE

The analysis of multiconductor edge-coupled systems has been extensively investigated [19]–[25]. The introduced methods are based on the use of the capacitance, immittance, or S-parameter matrix of the system. Recently, the multilayer broadside-coupled structures have received an increased interest due to the new multilayer techniques adopted by the modern design technology such as low-temperature co-fired ceramics and laminated multichip modules. The broadside-coupled microstrip-slot-microstrip structures were well investigated for the case of two coupled layers [18], [27], [28]. Here, a closed-form solution is presented for the analysis of the broadside-coupled three-layer structure, which uses a microstrip-slot-microstrip configuration. The proposed method is used to design arbitrary three-way power dividers.

The configuration of the proposed multilayer three-way power divider, which uses elliptical-shaped patches, is shown in Fig. 1. It consists of five conductive layers interleaved by four dielectrics. The input and one of the output ports (port 4), which are stripline ports, are located at the mid layer of the structure, while the other two microstrip output ports (ports 2 and 3) are at the top and bottom layers. The ground plane, which also includes the coupling slots, is at the second and fourth layers of the circuit. There are two additional ports, which are isolated from the input port, and thus, they have no power output. They are terminated in matched loads to absorb any reflected signal from the output ports, and hence, improve their isolation.

It is worthwhile to mention that the elliptical shape was chosen for the coupled structures shown in Fig. 1 because it provides some sort of a tapered coupled configuration. Hence, it gives an almost constant coupling factor across the UWB, which results in a constant power division across that band.

The structure presented in Fig. 1 can be fully analyzed using the three fundamental modes of operation: even–even, odd–odd, and odd–even. Distribution of the electric field lines between the three broadside-coupled layers for the three fundamental modes is shown in Fig. 2. The excitations needed to generate the three modes are shown in Fig. 2. They are considered according to the definition given in [19]–[24]. For the even–even mode, the three layers are excited in-phase, whereas in the odd–odd mode, the top and bottom layers are out-of-phase with respect to the mid layer. In the odd–even mode, the top and bottom layers are out-of-phase with each other and the mid layer is at zero potential. It is worth mentioning that some authors used the terms odd and even modes with symmetrical and asymmetrical coupled lines [29], [30], whereas other authors replaced it for the case of asymmetrical coupled lines with (k, k, k modes) [19], [24] or (k, k, k modes) [31].

The structure displayed in Fig. 1 has six ports. Due to symmetry, the performance of the three ports at the left of the structure, i.e., input port 1 and output ports (2 and 3), is similar to their counterparts, i.e., output port 4 and the two matched ports. According to the characteristics of the backward directional couplers, the input port and the two matched ports shown in Fig. 1 are isolated. Therefore, the analysis that follows concentrates only on the calculation of the coupling between the input port and the two output ports 2 and 3. The power output from port
is the mutual capacitance per unit length and refers to the mode, and its value is the effective.

is the phase velocity of the and the ground, whereas analyze performance of the three-way device. In Fig. 3, capacitances of the lines and the phase velocity on the lines be completely determined from the effective per unit length pansion, the electrical characteristics of the coupled lines can be calculated depending on the value of the input power (d) odd–even mode.

4 can be calculated depending on the value of the input power and the two coupled output powers.

Assuming a quasi-transverse electromagnetic propagation, the electrical characteristics of the coupled lines can be completely determined from the effective per unit length capacitances of the lines and the phase velocity on the lines [1]. Therefore, the structures shown in Fig. 3 can be used to analyze performance of the three-way device. In Fig. 3, represents the capacitance per unit length between the line and the ground, whereas is the mutual capacitance per unit length between the lines and .

For each of the three modes of propagation, the capacitance for each of the three lines can be determined from Fig. 3(b)-(d).

For the even–even mode,

\[ \zeta_{10e} = \zeta_1 + 2 \zeta_{12} + 2 \zeta_{13} \]

\[ \zeta_{20e} = \zeta_2 + 2 \zeta_{12} \]

\[ \zeta_{30e} = \zeta_3 + 2 \zeta_{13} \] (2)

For the odd–even mode,

\[ \zeta_{10o} \rightarrow \infty \]

\[ \zeta_{20e} = \zeta_2 + \zeta_{12} + 2 \zeta_{23} \]

\[ \zeta_{30e} = \zeta_3 + \zeta_{13} + 2 \zeta_{23} \] (3)

The characteristic impedance of each of the three lines at any of the three modes can be found using the relation [1]

\[ Z_{ijk} = \frac{1}{(v_{ijk})} \zeta_{ijk} \] (4)

where subscript refers to the line number and refers to the mode, is the phase velocity of the mode, and its value can be calculated from the relation [1]

\[ v_{ijk} = \frac{c}{\sqrt{\varepsilon_{ijk}}} \] (5)

where is velocity of light in free space and is the effective dielectric constant of the medium of propagation at the mode . For the three-way device presented in this paper, the broadside coupling between the three layers occurs almost entirely within the substrate. Therefore, the effective dielectric constants for the three modes can be considered equal, and each of them is equal to the dielectric constant of the substrate . Hence, the phase velocity for any mode is

\[ v = \frac{c}{\sqrt{\varepsilon_r}}. \] (6)

For the configuration of the three-way divider proposed in this paper and shown in Fig. 1, the input port is at the mid layer. This means that the odd–even mode, which occurs when the mid layer is at zero potential, has no effect on the value of the coupling between the input port at the mid layer (port 1 in Figs. 1 and 3) and the two output ports (ports 2 and 3 in Figs. 1 and 3), which are located at the top and bottom layer. Collier and El-Deeb [23] used the same assumption when they analyzed the case of a three parallel coupled lines and the input was connected to the center line. On the other hand, the odd–even mode defines the coupling factor between the two coupled output ports (ports 2 and 3).

In order to get a closed-form solution for a general three-layer broadside-coupled case, the structure assumed in this paper is asymmetrical. By extending the method presented by Crystal [29] and used to analyze the coupling between asymmetrical two edge-coupled lines, it is possible to find the coupling factor between the mid layer, where the input port is, and the top layer (F21) and between the mid layer and the bottom layer (F31).

\[ F_{21} = \frac{Z_{12e}Z_{20o}}{Z_{21e}Z_{20o}} \sqrt{(\zeta_{12e} + \zeta_{12o})(\zeta_{20e} + \zeta_{20o})} \] (7)

\[ F_{31} = \frac{Z_{32e}Z_{30o}}{Z_{31e}Z_{30o}} \sqrt{(\zeta_{32e} + \zeta_{32o})(\zeta_{30e} + \zeta_{30o})}. \] (8)
The coupling factors as a function of the capacitances can be obtained by substituting from (4) and (6) into (7) and (8) as follows:

\[
\zeta' F_{21} = \frac{\zeta_{2oo} - \zeta_{2ce}}{\sqrt{\left(\zeta_{1oo} + \zeta_{1ce}\right)\left(\zeta_{2oo} + \zeta_{2ce}\right)}} \\
\zeta' F_{31} = \frac{\zeta_{3oo} - \zeta_{3ce}}{\sqrt{\left(\zeta_{1oo} + \zeta_{1ce}\right)\left(\zeta_{3oo} + \zeta_{3ce}\right)}}. 
\]

(9)  
(10) 

It is possible to use a similar analysis to the one in [1], which was used for two coupled lines, to show that in order to design a divider with infinite directivity and a perfect matching at the input/output ports, which have characteristic impedance \(Z_\tau (= \sqrt{\mu \varepsilon})\), then

\[
\bar{Z}_0 = \sqrt{\bar{Z}_{1oo}\bar{Z}_{3oo}} = \sqrt{\bar{Z}_{2oo}\bar{Z}_{2oo}} = \sqrt{\bar{Z}_{1oo}\bar{Z}_{2oo}}. \tag{11}
\]

In deriving (11), it was assumed that the coupling between the mid layer and the top and bottom layers is defined by the even–even and odd–odd modes. Substituting from (11) in (7) and (8) results in

\[
\zeta' F_{21} = \frac{\zeta_{2oo} - \zeta_{2ce}}{\sqrt{\left(\zeta_{1oo} + \zeta_{1ce}\right)\left(\zeta_{2oo} + \zeta_{2ce}\right)}} \\
\zeta' F_{31} = \frac{\zeta_{3oo} - \zeta_{3ce}}{\sqrt{\left(\zeta_{1oo} + \zeta_{1ce}\right)\left(\zeta_{3oo} + \zeta_{3ce}\right)}}. \tag{12} \tag{13}
\]

Note that for the case of a traditional symmetrical two-line directional coupler, line 3 does not exist, whereas line 1 is similar to 2. Therefore, the coupling factor from (12) is \(\zeta' F_{23} = \frac{\left(\zeta_2 - \zeta_0\right)}{\left(\zeta_2 + \zeta_0\right)}\), which is the well-known equation for the ordinary two-line directional coupler [1].

If it is required to find the coupling factor between the top and bottom layer \(\zeta' F_{23}\), which is expected to be very small and negligible, an analysis similar to the one used for the coupling factors between the mid layer and the other two layers can be used. The final result is shown here, which is

\[
\zeta' F_{23} = \frac{\zeta_{2ce} - \zeta_{2ce} + \zeta_{2oo} + \zeta_{2oo}}{2} \left\{ \frac{\zeta_{2ce} + \zeta_{2ce} + \zeta_{2oo} + \zeta_{2oo}}{2} \right\} \left\{ \frac{\zeta_{3ce} + \zeta_{3ce} + \zeta_{3oo} + \zeta_{3oo}}{2} \right\} \\
= \frac{\zeta_{2ce} - \zeta_{2ce} + \zeta_{2oo} + \zeta_{2oo}}{2} \left\{ \frac{\zeta_{3ce} + \zeta_{3ce} + \zeta_{3oo} + \zeta_{3oo}}{2} \right\} \left\{ \frac{\zeta_{2ce} + \zeta_{2ce} + \zeta_{2oo} + \zeta_{2oo}}{2} \right\}. \tag{14}
\]

The equality is used in the above equation because

\[
\frac{\zeta_{2ce} - \zeta_{2ce} + \zeta_{2oo} + \zeta_{2oo}}{2} = \frac{\zeta_{3ce} - \zeta_{3ce} + \zeta_{3oo} + \zeta_{3oo}}{2} = 2 \zeta'_{23}. \tag{15}
\]

This can be verified from (1)–(3). For the device under consideration, a perfect isolation is required between the output ports. Therefore, the factor \(\zeta' F_{23}\) should equal to zero. From (14) and (15), this means that \(\zeta'_{23} \rightarrow 0\).

III. DESIGN OF ARBITRARY THREE-WAY POWER DIVIDERS

Assume that it is required to design a three-way power divider with a power ratio of \(\alpha : \beta : \gamma\) at the output ports. The required coupling factor between the mid layer, which is connected to the input port, and the top layer (\(\zeta' F_{21}\)), and between the mid layer and the bottom layer (\(\zeta' F_{31}\)) should be chosen such that

\[
\zeta' F_{21} = \frac{\alpha}{\sqrt{\left(\alpha + \beta + \gamma\right)\left(\alpha + \beta + \gamma\right)}} \quad \zeta' F_{31} = \frac{\gamma}{\sqrt{\left(\alpha + \beta + \gamma\right)\left(\alpha + \beta + \gamma\right)}}. \tag{16}
\]

The output power from the direct output port (port 4 in Fig. 1) is \(1 - \zeta' F_{21}^2 - \zeta' F_{31}^2 = \alpha/(\alpha + \beta + \gamma)\).

The dimensions of the elliptical coupled microstrips and slots offering the required coupling factors can be determined by extending the quasi-static approach presented in [18] to the case of a multilayer coupler. Using that approach with the help of the conformal mapping technique [26], the capacitances shown in Fig. 3 can be calculated as a function of the coupled structure’s dimensions. Using the results obtained for the capacitances, the impedances in (7) and (8) or (12) and (13) can be calculated. The final equations are

\[
\tilde{Z}_{1oo} = \frac{\varepsilon_r}{\varepsilon_f} \left[ \frac{K_v(l_1)}{K_v(l_2)} \right]^{-1} \\
\tilde{Z}_{2oo} = \frac{\varepsilon_r}{\varepsilon_f} \left[ \frac{K_v(l_3)}{K_v(l_2)} \right]^{-1} \\
\tilde{Z}_{3oo} = \frac{\varepsilon_r}{\varepsilon_f} \left[ \frac{K_v(l_4)}{K_v(l_2)} \right]^{-1} \tag{17} \tag{18} \tag{19}
\]

where \(\varepsilon_r\) is the dielectric constant of the substrate, \(K_v(l)\) is the first kind elliptical integral, and \(K_v(l_1) = K_v(\sqrt{1-l^2})\). \(l_1\), \(l_2\), \(l_3\), and \(l_4\) are the major diameters of the top, mid, and bottom coupled layers, respectively. The parameters \(l_2\) in (17)–(19) are equal to

\[
l_1 = \left[ \sin^2 \left( \frac{\pi^2 l_1 d_{dib}}{4h_1 \lambda_c} \right) + \cos^2 \left( \frac{\pi^2 l_1 d_{dib}}{4h_1 \lambda_c} \right) \right] \tag{20}
\]

\[
l_2 = \left[ \sin^2 \left( \frac{\pi^2 l_2 d_{dib}}{4h_1 \lambda_c} \right) + \cos^2 \left( \frac{\pi^2 l_2 d_{dib}}{4h_1 \lambda_c} \right) \right] \tag{21}
\]

\[
l_3 = \left[ \sin^2 \left( \frac{\pi^2 l_3 d_{dib}}{4h_1 \lambda_c} \right) + \cos^2 \left( \frac{\pi^2 l_3 d_{dib}}{4h_1 \lambda_c} \right) \right] \tag{22}
\]
where \( h \) is the thickness of the one layer substrate, \( \lambda_c \) is the wavelength in the propagation medium calculated at the center frequency of operation, and \( P_{st} \) and \( P_{sh} \) are the major diameters of the upper and lower slots, respectively. Lengths of the coupled layers depend mainly on the wavelength at the center frequency of operation. The method presented in [18] can be adopted to calculate the length of the top \( (l_t) \), mid \( (l_m) \), and bottom \( (l_b) \) coupled layer, and the upper \( (l_{st}) \) and lower \( (l_{sh}) \) slot as in the following equations:

\[
l_t = \frac{\lambda_c}{4} \left[ 1 - \left( \frac{\pi \Gamma_2}{2 \lambda_c} \right)^2 \right]^{-1}
\]

\[
l_m = \frac{\lambda_c}{4} \left[ 1 - \left( \frac{\pi \Gamma_4}{2 \lambda_c} \right)^2 \right]^{-1}
\]

\[
l_b = \frac{\lambda_c}{4} \left[ 1 - \left( \frac{\pi \Gamma_6}{2 \lambda_c} \right)^2 \right]^{-1}
\]

\[
l_{st} = \frac{\lambda_c}{4} \left[ 1 - \left( \frac{\pi \Gamma_{st}}{2 \lambda_c} \right)^2 \right]^{-1}
\]

\[
l_{sh} = \frac{\lambda_c}{4} \left[ 1 - \left( \frac{\pi \Gamma_{sh}}{2 \lambda_c} \right)^2 \right]^{-1}
\]

The design parameters used in (17)–(24) are shown in Fig. 1. Inspection of (24) reveals that length of the coupled structure is \( \lambda_c/4 \) when the required diameter approaches zero. For any other value, the length is larger than \( \lambda_c/4 \). Parametric and optimization analysis of the above equations using structures of different coupling factors indicate that the required length approaches a certain maximum value (\( \approx 1.5 \lambda_c/2 \)) when the diameter approaches \( \lambda_c/4 \), after which, any increase in the diameter has no significant effect on the optimum value for the length. Therefore, in order to generalize the use of (24), it is possible to suggest the following condition on it; (24) can be used when the diameter is less than \( \lambda_c/4 \). If it is larger than \( \lambda_c/4 \), then the length is equal to \( 1.5 \lambda_c/2 \). The above condition is based on the minimum mean-square-error fitting of the optimization results.

The required values for width of the input and output stripline \( (w_i, w_o) \), and microstrip ports \( (w_{i+m}, w_{o+m}) \) to give 50-\( \Omega \) characteristic impedance can be determined using the well-known stripline/microstrip equations [1].

Using the presented design method, dimensions of the 1 : 1 : 1, 2 : 1 : 1, and 4 : 2 : 1 three-way dividers were calculated and then optimized using Ansoft’s HFSSv10 software. The final dimensions (in millimeters) are shown in Table I. It was found that the optimized values of the design parameters (except \( l_{i+m} \)) are less than 5% different from those obtained by the described design method. The optimized value of \( l_{i+m} \) was found to be less than the calculated value by around 10%.

### Table I

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Dev. 1</th>
<th>Dev. 2</th>
<th>Dev. 3</th>
</tr>
</thead>
<tbody>
<tr>
<td>( l_t )</td>
<td>4.8</td>
<td>4.2</td>
<td>3.65</td>
</tr>
<tr>
<td>( l_m )</td>
<td>4.4</td>
<td>2.7</td>
<td>3.32</td>
</tr>
<tr>
<td>( l_b )</td>
<td>4.8</td>
<td>4.2</td>
<td>3.32</td>
</tr>
<tr>
<td>( l_{st} )</td>
<td>1</td>
<td>1.5</td>
<td>1.5</td>
</tr>
<tr>
<td>( l_{sh} )</td>
<td>1.5</td>
<td>1.5</td>
<td>1.5</td>
</tr>
<tr>
<td>( l_{i+m} )</td>
<td>0.8</td>
<td>0.8</td>
<td>0.8</td>
</tr>
</tbody>
</table>

Fig. 4. Manufactured 4 : 2 : 1 three-way power divider.

### IV. RESULTS

The validity of the presented design method was tested by building 1 : 1 : 1, 2 : 1 : 1, and 4 : 2 : 1 three-way power dividers aimed at the operation in the UWB 3.1–10.6-GHz range. Rogers RO4003C (with \( \varepsilon_r = 3.58, h_i = 1.524 \text{ mm} \), and loss tangent = \( 0.00127 \)) was used as a substrate. A photograph of one of the developed devices (with the ratio 4 : 2 : 1) is shown in Fig. 4. The developed devices have a compact size with an overall dimension of 20 mm × 30 mm. The manufactured power dividers were tested via simulations and measurements. The simulations were performed using Ansoft’s HFSSv10 commercial software,
Fig. 5. Measured and simulated performance of the 1:1:1 three-way divider. (a) Insertion loss and return loss. (b) Isolation.

whereas the measurements were done using a vector network analyzer.

The simulated and measured performance of the 1:1:1 divider are shown in Fig. 5. The developed device exhibits an insertion loss at the three output ports equal to 4.77 dB ± 1 dB across the 3.1–10.6-GHz band revealing an UWB performance. The return loss for the input/output ports and the isolation between the output ports of the device are better than 17 dB across the 3.1–10.6-GHz band. Note that because of symmetry, $S_{11} = S_{14}$, and because $F_{21} = F_{31}$ and due to symmetry, then $S_{22} = S_{33}$, $S_{24} = S_{34}$, and $S_{21} = S_{31}$.

The simulated and measured results for the 2:1:1 divider are presented in Fig. 6. The designed device shows an insertion loss of 3 dB ± 1 dB for the direct output (port 4), and 6 dB ± 1 dB for the two coupled outputs across the 3.1–10.6-GHz band indicating a 2:1:1 power division. The return loss for the input/output ports is better than 18 dB, whereas the isolation between the output ports of the device is better than 14 dB across the UWB. Note that because of symmetry, $S_{14} = S_{11}$, and because $F_{21} = F_{31}$ and due to symmetry, then $S_{22} = S_{33}$, $S_{24} = S_{34}$, and $S_{21} = S_{31}$.

The simulated and measured results for the 4:2:1 divider are depicted in Fig. 7. The manufactured device shows an insertion loss of 2.5 dB ± 1 dB for the direct output (port 4), 5.5 dB ± 1 dB for the top coupled output (port 2), and 8.5 dB ± 1 dB for the bottom coupled output (port 3) across the 3.1–10.6-GHz band revealing a 4:2:1 power division. The return loss for the input/output ports is better than 13 dB, whereas the isolation between the output ports of the device is better than 15 dB across the UWB.

Concerning the phase performance of the developed devices, the simulated and measured results indicated that the output signals from the coupled ports (2 and 3) are in phase, whereas the output signal from the direct port (port 4) has a 90° phase shift.
with respect to any of the coupled ports. This phase performance can be explained by referring to the fact that the building block of the proposed three-way divider is a quadrature coupler [18]. The simulated and measured performance of one of the developed dividers (the 4:2:1 power divider) is shown in Fig. 8. The phase difference between the output signals from the coupled ports is equal to zero according to the simulation, whereas it is equal to \( 3^\circ \pm 3^\circ \) according to the measured results. The phase difference between each of the coupled ports and the direct port is equal to \( 95^\circ \pm 5^\circ \) according to the simulation and \( 98^\circ \pm 8^\circ \) according to the measurements across the 2–11-GHz band. A better performance was noted for the 1:1:1 and 2:1:1 three-way power dividers, which have a compact size. The simulated and measured results of the manufactured devices have shown an UWB performance concerning the insertion loss, return loss, and isolation.

V. CONCLUSION

A method to design arbitrary three-way power dividers with UWB performance has been presented. The proposed devices utilize a broadside-coupled structure, which has three coupled layers. The design method exploits the three fundamental modes of propagation (even–even, odd–odd, and odd–even) and the conformal mapping technique to find the coupling factors between the different layers. The method has been used to design 1:1:1, 2:1:1, and 4:2:1 three-way power dividers, which have a compact size.

The analysis presented in this paper should find a good interest from the industry and the academy working in the emerging multilayer technology for UWB applications.

The multilayer three-way dividers introduced in this paper are especially suitable to the implementation in the modern multilayer structures such as the laminated multichip modules and the low-temperature co-fired ceramics. In such structures, the broadside coupling is much preferred from a reproducibility and loss perspective.

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Amin M. Abbosh received the M.Sc. degree in communications systems and Ph.D. degree in microwave engineering from Mosul University, Mosul, Iraq, in 1991 and 1996, respectively. Until 2003, he was a Head of the Computer and Information Engineering Department, Mosul University. In 2004, he joined Griffith University and then the School of Information Technology and Electrical Engineering, The University of Queensland, Brisbane, St. Lucia, Qld., Australia, as a Research Fellow. His research interests include antennas, radio wave propagation, microwave devices, and UWB wireless systems.
BROADBAND MULTILAYER IMPHASE POWER DIVIDER FOR C-BAND APPLICATIONS

A.M. Abbosh

A broadband inphase power divider for C-band applications is presented. The proposed divider utilises a broadband microstrip/slotline coupling, which is especially suitable for multilayer technology, where the output ports are located at different layers. Simulated and measured results show that the proposed device has an insertion loss of less than 0.2 dB and an isolation of more than 10 dB across the C-band (4–8 GHz).

Introduction: Power dividers are widely used in antenna arrays, power amplifiers, mixers, phase shifters and vector modulators. Wilkinson and the hybrid ring, also known as the Gysel power divider, are the most famous power dividers [1, 2]. The exploding growth of wireless-communication systems has led to increased demand for the multilayer integration technology, such as low temperature co-fired ceramic and laminated multi-chip modules. The power divider to be used in these technologies should have its output ports distributed in different layers [3]. The Wilkinson and Gysel power dividers are uniplanar devices. In order to use them in the multilayer technology, vertical transitions are to be used. However, the use of the additional transitions increases the insertion loss of the power dividers and limits their band of operation. In this Letter, the configuration of a compact multilayer power divider with broadband performance is presented. It utilises the broadside coupled microstrip/slot structure, which is compatible with the modern multilayer technology.

Design: The configuration of the proposed multilayer inphase power divider is shown in Fig. 1. It consists of three conductor layers interleaved by two dielectrics. The input port and one of the output ports are located at one layer of the structure, while the other output port is located at a different layer. The ground plane is at the mid layer of the structure. There is a slot in the shape of a narrow rectangle ending with two circles in the ground plane. This slot is responsible for guiding the wave from the input port to the two output ports. The other small circular slot at the ground plane, shown in Fig. 1b, is used as a passage for the isolation resistor R which is connected between the two output microstrip lines.

Fig. 1 Configuration of proposed power divider
a Top layer
b Mid layer
c Bottom layer
d Whole configuration

The divider utilises a T-junction formed by a slot line and two microstrip lines. To efficiently couple a signal from the slot line to the microstrip, the end of the slot line needs to be compensated with an inductive element. Here, it is chosen in the form of a circular slot. To convert the input port from the slot type to the microstrip type, a wideband slot to microstrip transition needs to be employed [4]. As seen in Fig. 1, the chosen transition is formed by two complementary structures. One includes a microstrip line terminated with a capacitive circular disk and the other one is a slot line terminated with an inductive circular slot. The exact location of these circles is depicted in Fig. 1d.

To improve the isolation between the two output ports, a resistor R is connected between the two microstrip lines which are connected to the two output ports, see Fig. 1. In the normal operation of the divider, and owing to symmetry, the two signals at ports 2 and 3 are equal in amplitude and in phase. Therefore, the two terminals of the resistor R are at the same potential and no current flows via this resistor and the device is lossless. If a mismatch occurs at any of the output ports, there is a reflected signal at that port, and therefore the two terminals of the resistor are in different potential. Hence, if the resistor’s value R is chosen properly, most of the reflected signal is dissipated in R and the two output ports are kept isolated. A parametric analysis using the software HFSSv10 indicated that R should be around \( \frac{1}{2} \times Z_0 \) for the best isolation, where \( Z_0 \) is the characteristic impedance of the output ports.

Having established the principles of operation of the power divider, a simple procedure can be applied to its design. The width of the input and output microstrip ports is determined assuming a (50 Ω) characteristic impedance. The length of the slot line at the ground plane (l0) is chosen to be a quarter of the effective wavelength at the centre frequency of operation (6 GHz).

To achieve a high return loss at the microstrip ports, the slot width (S) should be chosen to give impedance close to 50 Ω as seen from the microstrip side. At the same time, the slot width should not be too narrow, to avoid problems with manufacturing errors. According to [4], impedance of the slot as seen from the microstrip line (Zs) is equal to \((N^2 Z_0)\) where N and the slot impedance Zs depend on the slot width and the substrate characteristics as given in [4, 5]. In the present design, the slot width is chosen to give Zs = 100 Ω and N = 1/(\( \sqrt{2} \)) so that the impedance of the slot as seen from the microstrip line is \((1/\sqrt{2})^2 \times 100 = 50 \ Ω\).

Although the slot line is chosen to have characteristic impedance equal to 50 Ω, the effective input impedance of the structure as seen at the input port is less than 50 Ω because the slot line is effectively loaded by two parallel output ports. Hence, to improve the matching at the input port, a tapered microstrip line, which has a width equal to w0 at the input port and w at the microstrip/slot transition, is used to connect the input port to the coupled microstrip/slot line region as shown in Fig. 1a.

The radius of the input microstrip circle (r0), and the output microstrip circle (r) terminating the input/output microstrip lines are chosen to be between one and two times of the microstrip width, while the radius of the slot circle (r1) is usually chosen to be less than that since it is used to terminate a slot line which has a narrow width compared with the width of the microstrip lines. This choice conforms well to the guidelines for designing wideband microstrip/slot line transitions [4].

Results: The validity of the presented design method was tested by building a power divider aimed at operation in the C-band (4–8 GHz). Rogers RO3010 Dielectric (with \( \varepsilon_r = 3.38, h = 0.508 \) mm and loss tangent = 0.0027) was used for the divider’s development. Using the proposed design procedure and with the help of the optimisation capability of the software Ansoft HFSSv10, parameters of the divider were found to be \( r_0 = 1.9 \) mm, \( r = 1.5 \) mm, \( r_1 = 0.8 \) mm, \( w_0 = 1.15 \) mm, \( w = 1.46 \) mm, \( S = 0.11 \) mm, \( l_0 = 6.9 \) mm, and \( R = 63.4 \) Ω. The manufactured power divider has overall dimensions of 25 × 30 mm. The device was tested via simulation and measurement. Simulation was performed using the commercial software Ansoft HFSSv10, whereas measurement was done using a vector network analyser. Sub-miniature A (SMA) connectors were used to connect the manufactured device to the measuring tool.

The insertion loss, return loss and isolation of the developed device are shown in Fig. 2. The insertion loss is equal to 3.4 ± 0.2 dB across the band 4 to 8 GHz revealing a broadband performance with less than ±0.2 dB amplitude imbalance between the output signals. The return loss at the three ports of the device is better than 10 dB across the C-band, whereas the isolation between the two output ports is about 10 dB across most of the C-band. Concerning the phase performance of the device, the measured and simulated results shown in Fig. 3.
indicate that the two output signals are in phase with less than $\pm 2^\circ$ phase difference.

**Fig. 2** Simulated and measured S-parameters of power divider

**Fig. 3** Phase performance of power divider

**Conclusion:** A compact multilayer inphase power divider with broad-band behaviour is presented. The proposed divider utilises broadside coupling via a multilayer microstrip/slot configuration. Simulated and measured results of the developed device show equal power division with $3.4 \pm 0.2$ dB insertion loss, good return loss and isolation over the C-band.

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**References**

the other hand, it is also easy to design the waveguide-type
directional coupler integrated into the thin dielectric substrate
when the specifications are given.

4. FREQUENCY LIMITATION (FORBIDDEN FREQUENCIES)
OF THE FORMULA
It must be noted that, above formulas cannot be used in any
situation, the restriction to these formulas is now discussed. From
Eqs. (3) and (4), the coupling and isolation voltages of the slot
cannot be zero in the working frequency band, which implies that
these formulas cannot be used at certain frequencies (forbidden
frequencies exist), i.e. the following equations must be satisfied.

\[ -q \left( \frac{\pi}{a} \right)^2 + \frac{p^2}{4p - 1} \left( \frac{\lambda_c \cos \pi}{a} \right)^2 - q^2 \left( \frac{\lambda_c \sin \pi}{a} \right)^2 \neq 0 \]

\[ -q \left( \frac{\pi}{a} \right)^2 + \frac{p^2}{4p - 1} \left( \frac{\lambda_c \cos \pi}{a} \right)^2 - q^2 \left( \frac{\lambda_c \sin \pi}{a} \right)^2 \neq 0 \]

The forbidden frequencies can be found as

\[ f = \frac{c}{2a} \left( \frac{\left( \frac{p}{q} \right)^2 \cos \left( \frac{\pi}{a} \right) - 1}{4p - 1} + 1 \right) \]

Or

\[ f = \frac{c}{2a} \left( \frac{\left( \frac{p}{q} \right)^2 \cos \left( \frac{\pi}{a} \right) - 1}{4p - 1} - 1 \right) \]

This indicates the derived formulas are not valid when the oper-
ational frequency is located at the forbidden frequency.

5. CONCLUSION
The waveguide-type directional couplers integrated into dielectric
substrate are investigated in this article. Based on traditional cou-
lpling theory of directional coupler, the formulas to calculate the
coupling and isolation of the SIW directional couplers are derived,
in cases of single row slots and double row slots, respectively.
Numerical simulation results have been compared with the calcu-
lations, good agreements have been observed, demonstrating the
good accuracy of these formulas. With these formulas it is conve-
nient to calculate the performances of a ready-made SIW direc-
tional coupler or to design an engineering SIW directional coupler
in case of known specifications. We may indicate that these for-
mulas can be valid except in case coupling slots with large size.
With these in mind, our formulas can find potential applications in
engineering work of LTCC or multilayer PCB technologies.

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1. INTRODUCTION

Power dividers play an important role in the design of microwave circuits. They are widely used in antenna arrays, power amplifiers, mixers, phase shifters, and vector modulators [1]. The most widely used power divider is the Wilkinson divider [2]. It has completely matched output ports with sufficiently high isolation between them. Moreover, it offers equal phase characteristics at each of its output ports. This device is also potentially without loss provided that no reflected power from output ports enters into it. However, in its original configuration, the Wilkinson power divider has a narrow bandwidth. Several modifications have been proposed later to either increase its bandwidth or reduce its size [3–6]. The techniques used to reduce the size of the Wilkinson divider ended up in a very narrowband performance [5, 6].

The other widely used inphase power divider, the hybrid-ring [7], also known as the Gysel power divider, usually assumes the shape of a ring impedance transformer. The Gysel power divider can be regarded as a Wilkinson power divider terminated with a transmission stub load. Because the Gysel divider is a narrowband device, several methods have been proposed to increase its bandwidth [8, 9]. However, it still has less than 50% fractional bandwidth, which prevents its use in ultra wideband (UWB) applications.

The exploding growth of wireless communication systems has led to an increasing demand for the multilayer integration technology, such as the low temperature co-fired ceramic and the laminated multichip modules, and to the use of these technologies in the design of UWB systems. The power divider is among the passive components needed for use in these modern technologies. The power divider to be used in these technologies should have its three ports distributed in different layers. The Wilkinson and Gysel power dividers are uniplanar devices. Hence, to use them in the multilayer technology, vertical transitions are to be used. However, the use of the additional transitions increases the insertion loss of the power dividers and limits their band of operation.

In this paper, the configuration of a compact multilayer power divider with UWB performance is presented. It utilizes the broadside-coupled structure that is compatible with the modern multilayer technology. Simple rules are derived to design the device and estimate its performance. The simulated and measured results confirm the good performance of the proposed device across the band 3.1–10.6 GHz. The proposed divider exhibits $3.3 \pm 0.3$ dB insertion loss, about 20 dB return loss at its input port, and 10 dB isolation between the output ports across the ultra wide frequency band.

2. DESIGN

The configuration of the proposed multilayer power divider is shown in Figure 1. It consists of five conductor layers interleaved by three dielectrics. The input port is located at the mid layer of the structure, whereas the output ports are at the top and bottom layers. The ground plane, which also includes the coupling slot, is at the second and fourth layers of the structure. The microstrip-coupled patches and the slots are of elliptical shapes, similar to those used by the author to fabricate the UWB three-way power divider [10].

Assume that the device is designed to have a coupling equal to $C$ between the mid layer and the top and bottom patches, and that the input and output signals to/from the $i$-th port are $a_i$ and $b_i$, respectively. Depending on the odd-even modes analysis of the structure whose ports are depicted in Figure 2, the reflected signal at port 1 (the input port) and the output signals at port 2 and 3 (the output ports) can be calculated as follows [10–13]:

$$b_1 = \chi a_1 + \delta(a_4 + a_5)$$

$$b_2 = b_3 = \xi a_1 + \delta a_4$$

Figure 1 Figure configuration of the proposed power divider, (a) Top layer, (b) second layer (ground with coupling slot), (c) mid layer, (d) fourth layer (ground with coupling slot), (e) bottom layer, and (f) the whole configuration. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
For the structure under investigation, it is possible to show that

\[ b_4 = a_4 = \chi a_i \quad (3) \]

\[ b_5 = a_5 = b_6 = a_6 = \delta a_i \quad (4) \]

\[ \delta = \sqrt{1 - C^2 \cos(\beta l) + j \sin(\beta l)} \quad (5) \]

\[ \xi = \sqrt{1 - C^2 \cos(\beta l) + j \sin(\beta l)} \quad (6) \]

\[ \chi = \sqrt{1 - C^2 \cos(\beta l) + j \sin(\beta l)} \quad (7) \]

where \( l \) is the physical length of the coupled structure and \( \beta \) is the effective phase constant in the medium of the coupled structure.

For the structure under investigation, it is possible to show that

\[ \beta = \frac{\beta_e + \beta_o}{2} = \frac{2\pi}{\lambda} \sqrt{\varepsilon_r} \quad (8) \]

where \( \beta_e \) and \( \beta_o \) are the phase constants for the even and odd modes, respectively, \( \lambda \) is the free space wavelength, and \( \varepsilon_r \) is the dielectric constant of the substrate. In deriving (1–7), it was assumed that the output ports are perfectly matched. Because the ports 4, 5, and 6 are terminated in an open circuit then the return loss performance at the centre of the band is better.

3. RESULTS

The validity of the presented design method was tested by building a power divider aimed at the operation in the UWB range 3.1–10.6 GHz. Rogers RO4003C (with \( \varepsilon_r = 3.38, h = 0.508 \) mm, and loss tangent = 0.0027) was selected for the development of the divider. Using the proposed design method and with the help of the optimization capability of the software Ansoft HFSSv10, parameters of the divider were found to be as follows: \( D_m \) for the top and bottom layers = 5.1 mm, \( D_o \) for the mid layer = 3.3 mm, \( D_t = 6.7 \) mm, \( l = 7.4 \) mm, width of the stripline input port = 0.67 mm, and width of the microstrip output ports = 1.15 mm. Photo of the manufactured power divider is shown in Figure 4. It has a compact size with an overall dimension of 20 mm × 30 mm.

The designed power divider was tested via simulation and measurement. The simulation was performed using the commer-

Using \( C = 0.55 \), the even \( (Z_{eo}) \) and odd \( (Z_{oo}) \) mode characteristic impedances for each of the coupled patches are calculated using the following equations:

\[ Z_{eo} = Z_0 \sqrt{\frac{1 + C}{1 - C}} \quad (11) \]

\[ Z_{oo} = Z_0 \sqrt{\frac{3 - C}{1 + C}} \quad (12) \]

where \( Z_0 \) is the characteristic impedance of the input/output ports of the coupler.

Assuming that \( Z_0 = 50 \) \( \Omega \) and \( C = 0.55 \), then \( Z_{eo} = 92.8 \) \( \Omega \) and \( Z_{oo} = 26.94 \) \( \Omega \). The major diameters of the elliptical slot \( (D_e) \) and the coupled microstrip patches \( (D_m) \) offering the required even and odd mode characteristic impedances can be determined by using the quasi-static approach presented in [10]. The length of the slot and the coupled microstrip patches \( (l) \) are equal to quarter of the effective wavelength calculated at the center frequency, which is 6.85 GHz. The last step of the design is to calculate the width of the input and output stripline/microstrip ports. They are determined to give 50 \( \Omega \) characteristic impedance using the well known stripline/microstrip equations [1].

Figure 3  Variation of the calculated return loss and insertion loss with frequency. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
cial software Ansoft HFSSv10, whereas the measurement was done using a vector network analyzer.

The insertion loss of the two output ports, as shown in Figure 5, is equal to $3.3 \pm 0.3$ dB across the band 3.1–10.6 GHz revealing an UWB performance with less than $\pm 0.3$ dB amplitude imbalance between the output signals. The return loss for the input port of the device and the isolation between the output ports are also shown in Figure 5. The return loss at the input port is better than 20 dB across most of the UWB, whereas the isolation between the two output ports is about 10 dB across the band 3.1–10.6 GHz.

Concerning the phase performance of the device, the measured and simulated results shown in Figure 6 indicate that the two output signals are in phase with less than $\pm 2^\circ$ phase difference.

4. CONCLUSION

A multilayer inphase power divider with UWB behavior has been presented. It has a compact size with an overall dimension of 20 mm $\times$ 30 mm. The proposed divider utilizes broadside coupling via a multilayer microstrip/slot configuration. The simulated and measured results of the developed device have shown equal power division with $3.3 \pm 0.3$ dB insertion loss, good return loss, and isolation over the band 3.1–10.6 GHz.

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Figure 4 Photo of the manufactured power divider. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

Figure 5 The simulated and measured S-parameters of the power divider. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

Figure 6 The phase performance of the power divider. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
Ultra wideband inphase power divider for multilayer technology

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Abstract: A multilayer inphase power divider with an ultra wideband behaviour is presented. The proposed divider exploits broadside coupling via a multilayer microstrip/slot configuration. The design method utilised for the device is based on the conformal mapping techniques. The developed device has a compact size with an overall dimension of 20 mm x 30 mm. The simulated and measured results show that the proposed device has equal power division between the two output ports with <0.2 dB amplitude imbalance between them, better than 10 dB return loss and isolation and <2° phase difference between the two output signals across the frequency band 3.1–10.6 GHz.

1 Introduction

Power dividers play an important role in the design of microwave circuits. They are widely used in antenna arrays, power amplifiers, mixers, phase shifters and vector modulators [1]. The exploding growth of wireless communication systems has led to a huge increase in the demand for the multilayer integration technology, such as the low temperature co-fired ceramic (LTCC) and the laminated multi-chip modules (LMCM), and to the use of these new technologies in the design and fabrication of ultra wideband (UWB) systems. The power divider is among the passive components needed in these modern technologies and systems. The power divider to be used in LTCC or LMCM should have its three ports distributed in different layers.

The most widely used power divider is the Wilkinson divider [2]. It has completely matched output ports with sufficiently high isolation between them. Moreover, it offers equal-phase characteristics at each of its output ports. This device is also potentially lossless provided that no reflected power from output ports enters into it. However, in its original configuration, the Wilkinson power divider has a narrow bandwidth, which makes it inconvenient for UWB applications. Several modifications have been proposed later to either enhance its performance or reduce its size [3–10]. The techniques used to reduce the size of the Wilkinson divider ended up in a narrowband performance. Recently, the parallel strip lines have been used with a modified configuration of the Wilkinson power divider [11, 12]. The measured performance of the developed devices shows a performance which cannot cover the whole UWB.

The other widely used inphase power divider is the hybrid-ring [13], also known as the Gysel power divider, which usually takes the shape of a ring impedance transformer. The Gysel power divider can be regarded as a Wilkinson power divider terminated with a transmission stub load. Because the Gysel divider is a narrowband device, several methods have been proposed to increase its bandwidth [14, 15]. However, it still has <50% fractional bandwidth which prevents its use in UWB application.

In this paper, the configuration of a compact multilayer power divider with UWB performance is presented. It utilises the broadside-coupled configuration, where the three ports of the device are distributed among different layers. This makes the proposed divider compatible with the modern multilayer technology. The simulated and measured results confirm the good performance of the proposed device across the band 3.1–10.6 GHz. The proposed divider exhibits equal power division between the output ports with <0.2 dB amplitude imbalance and <2° phase difference. The return loss at its three ports and
the isolation between the output ports are better than 10 dB across the whole UWB.

2 Design

The configuration of the proposed multilayer power divider is shown in Fig. 1. The input port is located at the mid layer of the structure, whereas the output ports are at the top and bottom layers. The ground plane, which also includes the coupling slot, is at the second and fourth layers of the structure. The microstrip-coupled patches and the slots are of elliptical shapes. This shape behaves like a tapered transmission line which helps to achieve an almost constant value for the coupling factor across the UWB as proven in the results presented in [16]. The whole structure of the power divider is shown in Fig. 2.

An outline of the different ports of the device is shown in Fig. 3. There is a resistor \( R \) connected between the two output ports. In the normal operation of the divider, the input signal connected to port 1 is divided between the output ports (ports 2 and 3). Because of symmetry between the top and bottom layers, the two signals at ports 2 and 3 are equal in amplitude and phase. Therefore the two terminals of the resistor \( R \) are at the same potential, which means that no current flows via this resistor and the device is lossless and behaves as if the resistor does not exist. If a mismatch occurs at any of the output ports, there is a reflected signal at that port and the effect of the resistor appears. It is possible to compare between two cases: without and with the resistor \( R \). If there is no resistor, the reflected signal from the mismatched port, port 2, for example, is coupled to port 4. Because this port is terminated in an open circuit, the coupled power is reflected back and part of it is coupled to the two output ports. In conclusion, part of the reflected signal from port 2 emerges from port 3, which means a bad isolation between them. If the resistor \( R \) is connected between the two output ports, the two terminals of the resistor are in different potentials because of the existence of the reflected signal. Hence, if the resistor’s value \( R \) is chosen properly, most of the reflected signal is dissipated in \( R \) and the isolation between the two output ports is improved.

Assume that the device is designed to have a coupling factor equal to \( C \) between the mid layer and the top and bottom layers and that the input and output signals to/from the \( i \)th port are \( a_i \) and \( b_i \), respectively. Depending on the odd–even mode analyses of the structure, whose ports are depicted in Fig. 3, the reflected signal from port 1 (the input port) and the output signals from ports 2 and 3 (the output ports) can be calculated as follows [17, 18]

\[
\begin{align*}
\delta &= \frac{jC \sin(\beta l)}{\sqrt{1 - C^2 \cos(\beta l)} + j \sin(\beta l)} \\
\xi &= \frac{\sqrt{1 - C^2}}{\sqrt{1 - C^2 \cos(\beta l)} + j \sin(\beta l)} \\
\chi &= \frac{\sqrt{1 - C^2 - C^2 \sin^2(\beta l)}}{\sqrt{1 - C^2 \cos(\beta l)} + j \sin(\beta l)}
\end{align*}
\]

where \( l \) is the physical length of the coupled structure and \( \beta \)
the effective phase constant in the medium of the coupled structure.

For the structure under investigation, it is possible to show that
\[ \beta = (\beta_e + \beta_o)/2 = 2\pi\sqrt{\varepsilon_r/\lambda} \]
where \( \beta_e \) and \( \beta_o \) are the phase constants for the even and odd modes, respectively, \( \lambda \) the free space wavelength and \( \varepsilon_r \) the dielectric constant of the substrate. In deriving (1)–(7), it was assumed that the output ports are perfectly matched. The reflection coefficient at the ports 4–6 was assumed to be equal to 1 because these ports are terminated in an open circuit.

Assuming a lossless device, the return loss at the input port \( S_{11} \) and the insertion loss from the input to the output \( S_{21}, S_{31} \) can be computed from (1)–(7)

\[ S_{11} = \frac{1 - C^2 - 3C^2 \sin^2(\beta l)}{\sqrt{1 - 4C^4 \sin^2(\beta l)}} \]  \( (8) \)

\[ S_{21} = S_{31} = \frac{1 - |S_{11}|^2}{2} \]  \( (9) \)

It is possible to solve (8) and (9) in order to find the optimum value for the coupling factor, which results in the best possible performance of the proposed multilayer divider. If the length of the coupled structure \( l \) is chosen to be quarter of the effective wavelength at the design frequency, then the angle \( \beta l \) in (8) is equal to \( \pi/2 \). Using this value in (8) and (9) results in \( C = 0.5 \) for \( S_{11} = 0 \) and \( S_{21} = S_{31} = 0.707 \) (or 3 dB), which are the required values for an ideal performance. In order to make sure of the above results across the whole UWB, variation of the phase constants for the even and odd modes, respectively, for the top and bottom coupled layers.

The mid layer is connected to the input port and has nothing to do with the isolation resistor \( R \). The even- and odd-mode characteristic impedances of this layer \( Z_{1oe} \) and \( Z_{1oo} \) are chosen according to the following equation in order to achieve a perfect matching with the input port, whose characteristic impedance is \( Z_o \) [1]

\[ \sqrt{Z_{1oe}Z_{1oo}} = Z_o \]  \( (10) \)

Owing to the use of the isolation resistor \( R \) between the top and bottom layers, the perfect matching with the input/output ports requires that the mid layer has even- and odd-mode characteristic impedances, which are not essentially equal to those of the top and bottom layers. Therefore the structure, in this context, can be considered as asymmetrical and the method presented by Cristal [20] can be utilised to show that the coupling factor between the mid layer and any of the other layers is equal to

\[ C = \frac{Z_o}{\sqrt{Z_{2oe}Z_{2oo}}\sqrt{(Z_{1oe} + Z_{1oo})(Z_{1oe} + Z_{2oo})}} \]  \( (11) \)

It is to be noted that for the three-layer structure explained in this paper, the top and bottom layers are identical. The coupling between them is negligible, and thus it is possible to approximate those three coupled layers with a separate pair of two asymmetrical coupled layers [21]. Hence, the theory of a
A pair of coupled lines can be used to calculate the required odd- and even-impedances for the different coupled layers [21, 22].

The relation between the even- and odd-mode impedances of the different layers of the device and the physical dimension of the coupled structure can be found using the conformal mapping techniques [23] after assuming the quasi-static approach presented in [16, 22, 24] to the case of multilayer configuration. The major diameters of the elliptical slot (D) and the coupled microstrip patches at the mid layer (D1) and top/bottom layers (D2) offering the required even- and odd-mode characteristic impedances can be determined by using the following equations

\[ Z_{1oe} = \frac{30\pi K(k_1)}{\sqrt{\epsilon_r} K'(k_1)}, \quad Z_{1oo} = \frac{30\pi K'(k_2)}{\sqrt{\epsilon_r} K(k_2)} \] (13)

\[ Z_{2oe} = \frac{30\pi K(k_3)}{\sqrt{\epsilon_r} K'(k_3)}, \quad Z_{2oo} = \frac{30\pi K'(k_4)}{\sqrt{\epsilon_r} K(k_4)} \] (14)

\[ k_1 = \frac{\sinh^2(0.6D_1/b)}{\sinh^2(0.6D_1/b) + \cosh^2(0.6D_1/b)} \] (15)

\[ k_2 = \tanh\left(\frac{0.6D_1}{b}\right) \] (16)

\[ k_3 = \frac{\sinh^2(0.6D_2/b)}{\sinh^2(0.6D_2/b) + \cosh^2(0.6D_2/b)} \] (17)

\[ k_4 = \tanh\left(\frac{0.6D_2}{b}\right) \] (18)

where \( b \) is the thickness of the substrate, \( K(k) \) the first kind elliptical integral and \( K'(k) = K(\sqrt{1-k^2}) \).

The length of the coupled structure (\( l \)) is chosen to be equal to the quarter of the effective wavelength calculated at the design frequency, which is the centre frequency of the UWB, that is, 6.85 GHz.

There are several procedures which can be used to solve the design equations (10)–(18) and find the values of the design parameters. The procedure which is adopted in this paper can be summarised in the following steps

1. Assume a certain value for the odd impedance of the mid layer (\( Z_{1oo} \)). This value should be less than the value of the characteristic impedance of the input/output ports (50 \( \Omega \)).

2. Use (11) to find the even-mode impedance of the mid layer (\( Z_{1oe} \)).

3. Use (13)–(16) to find \( D_1 \) and \( D_2 \).

4. Use the known value of the coupling factor \( C \) and (12), (14), (17) and (18) to find \( D_2 \).

5. Use (10) to find \( R \).

6. If the calculated dimensions are reasonable and practical, that is, compact and easy to manufacture, then the design is complete; otherwise assume a new value for \( Z_{1oo} \) and repeat the steps 2–5.

3 Results and discussions

The validity of the presented design method was tested by building a power divider aimed at the operation in the UWB range. Rogers RO4003C (with \( \varepsilon_r = 3.38, \ b = 0.508 \) mm and loss tangent = 0.0027) was employed as a substrate. Assuming that \( Z_0 = 50 \) \( \Omega \), \( C = 0.55 \) and using the proposed design method, the values of the design parameters were found to be \( D_1 = 4.2 \) mm, \( D_2 = 3.1 \) mm, \( D_3 = 6.9 \) mm \( l = 6 \) mm, \( R = 52 \) \( \Omega \), width of the stripline input port = 0.68 mm and width of the microstrip output ports = 1.18 mm. With the help of the optimisation capability of the software Ansoft HFSSv10, the final values of the design parameters were found to be \( D_1 = 5 \) mm, \( D_2 = 2.8 \) mm, \( D_3 = 6.4 \) mm \( l = 7.6 \) mm, \( R = 51.5 \) \( \Omega \), width of the stripline input port = 0.67 mm and width of the microstrip output ports = 1.15 mm. A comparison between the calculated and optimised values reveals the capability of the proposed design procedure to find a good initial estimation for the required values of the design parameters. The manufactured power divider, which is shown in Fig. 5, has a compact size with an overall dimension of 20 mm \( \times \) 30 mm. The device was tested via simulations using HFSSv10, and measurements using a vector network analyzer.

The simulated and measured insertion loss of the two output ports, as shown in Fig. 6, indicates that the power is equally divided between the two output ports with \(<0.2 \) dB amplitude imbalance between them across the band 3.1–10.6 GHz revealing a UWB performance. The simulated insertion loss is 3.4 \( \pm 0.2 \) dB, whereas it is 3.6 \( \pm 0.2 \) dB in the measured results across the band.
3.1–10.6 GHz. At the centre of the band, the insertion loss is about 3.2 dB in the simulated results and 3.4 dB in the measured results. It is to be noted that the ideal value for the insertion loss of the equal division two-way power divider is 3 dB.

The return loss for the input and output ports of the device and the isolation between the output ports are also shown in Fig. 6. They are better than 10 dB across the whole UWB according to the simulated and measured results. The return loss at the three ports of the device and the isolation between the output ports are better than 13 dB across most of the UWB. There is a good agreement between the simulated and measured results depicted in Fig. 6.

Concerning the phase performance of the device, the measured and simulated results shown in Fig. 7 indicate that the proposed device is an inphase power divider, where the two output signals are in phase with $<2^\circ$ phase difference.

### 4 Conclusion

A compact multilayer inphase power divider with a UWB behaviour has been presented. The proposed divider utilises broadside coupling via a multilayer configuration. Assuming a quasi-static transverse electromagnetic approach, the conformal mapping technique was used to find the relation between the required performance and the physical dimension of the coupled structure. The simulated and measured results of the developed device have shown equal power division between the two output ports with $<0.2$ dB amplitude imbalance between them, good return loss and isolation and $<2^\circ$ phase difference between the output signals across the band 3.1–10.6 GHz.

### 5 References


negative mode to occur at 6.1 GHz, a length of patch in mushroom is readjusted. The resulting length and radius of via are determined to be 2.65 mm and 0.25 mm, respectively. As shown in Figure 4, the first negative mode occurs at 6.1 GHz. The Table 1 compares the capacitance values and sizes of conventional half wavelength gap coupled RH filter and mushroom CRLH filter. The CRLH filter using the first negative resonant mode has the reduced size by 41% over that of conventional half wavelength gap coupled RH filter. Note that the frequency responses of both RH and CRLH filter are almost the same even though they are not shown in this article.

Using the values in Table 1, the frequency responses of CRLH bandpass filter have been simulated by commercial EM (HFSS) and circuit simulators (Ansoft Designer) as shown in Figure 5. The results of both EM and circuit simulation shows good agreements.

However, some insertion loss of 1.5 dB in EM simulation is calculated due to material loss and radiation loss. Figure 6 compares the results of the measurement and EM simulation. The measured frequency response is up-shifted due to the fabrication tolerance. The insertion loss is also measured to be 3 dB, which is higher than that of simulation. Except for the frequency shift and the level of insertion loss, the overall frequency responses are in good agreements.

4. CONCLUSION
In this article, the high order \( (N = 2) \) CRLH bandpass filter is proposed using the first negative resonant mode. The design equation of CRLH filter is derived to design a high order CRLH filter. The resonators of proposed filter are implemented by two unit-cells of mushroom structure supporting the first negative resonant mode of CRLH TL. The size of CRLH filter is reduced by 41% when compared with that of conventional half wavelength gap coupled RH filter. The design processes are confirmed by measured and simulated results of CRLH bandpass filter.

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PLANAR ULTRA WIDEBAND INPHASE POWER DIVIDER

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ABSTRACT: A planar inphase power divider, which has ultra wideband performance, is presented. The proposed device utilizes a T-microstrip junction combined with an electromagnetic coupling between a
slotted ground plane and an elliptical patch at the centre of the T-junction. A resistor is also used to enhance the isolation between the output ports of the power divider. The simulated and measured return losses at the three ports of the device, isolation between its output ports, and phase stability of the output signals prove the ultra wideband performance of the presented device. © 2009 Wiley Periodicals, Inc.

1. INTRODUCTION

Power dividers are widely used in antenna feeds, balanced mixers, modulators, balanced amplifiers, phase shifters, automatic signal level control, signal monitoring, and many other applications.

The most widely known power dividers are the Wilkinson, hybrid ring, and T-junction [1–3]. Although several methods have been recently proposed to enhance the frequency band of operation for the Wilkinson and hybrid ring power dividers, their useful bandwidth is still less than the 110% fractional bandwidth required for the ultra wideband (UWB) applications [4, 5]. On the other hand, the T-junction power divider has a simple structure, but it has a poor isolation between its output ports [6].

In a recent development in designing broadband inphase power divider, the multilayer technology has been used to build compact devices [7, 8]. However, the measured performance has indicated either a 66% fractional bandwidth, which is inconvenient for UWB applications [7], or a limited isolation [8].

In this article, the simple T-junction power divider is modified to improve the return loss at the three ports of the device and enhance the isolation between the output ports over the UWB. The modification is achieved via the use of a controlled coupling between a slotted ground plane and an elliptical microstrip patch at the centre of the T-junction, which is located at the top layer of the structure. A resistor is connected across the slot at the ground plane in order to absorb any reflected signal at the output ports, thus improving the isolation between them. The design of the proposed device is achieved following a systematic approach. The simulated and measured performance of the device is presented to prove the validity of the described method.

2. DESIGN

The configuration of the proposed power divider is shown in Figure 1. The three ports of the divider are connected via an elliptical patch to form a T-junction and they are located at the top layer of the substrate. The ground plane is located at the bottom layer of the structure. A slot in the shape of a narrow rectangle is made in the middle of the ground plane directly under the elliptical patch of the top layer. To efficiently couple the signal between the slotline and the two output ports, the end of the slotline needs to be compensated with an inductive element. Here, it is chosen in the form of a circular slot. The coupling between the slot in the ground plane and the elliptical microstrip patch at the top layer is used to control performance of the device. Shape of the coupled microstrip patch was chosen to be elliptical because it represents a tapered shape, which enables a broadband performance [9].

Length of the coupling region (l) is chosen to be quarter of the effective wavelength \( \lambda_{ef} \) calculated at the centre of the UWB, that is, at 6.85 GHz. When calculating value of \( l \) and \( \lambda_{ef} \), it is assumed that the effective dielectric constant is equal to \((\varepsilon_r + 1)/2\), where \( \varepsilon_r \) is the dielectric constant of the substrate.

Figure 1 Configuration of the proposed power divider. (a) top layer highlighted, (b) bottom layer highlighted. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

An isolation resistor \( R \) is connected across the slotted ground plane. This resistor is used to absorb any power reflected from any of the output ports, hence improving the isolation between them.

To understand the operation of the proposed device, the even-and odd-mode analysis is used. The circuit that represents the proposed device is shown in Figure 2. Two voltage sources are considered at the two output ports to enable the even- and odd-mode analysis. The circuit of Figure 2 is drawn such that it is symmetrical with respect to its midplane. The characteristic impedance of all the three ports is considered to be \( Z_0 \). The impedance of the input port is divided into two parallel connected impedances each equal to \( 2Z_0 \).

In the even-mode, the two voltages at the ports 2 and 3 are equal in amplitude and phase. Thus, the two end points of the resistor \( R \) are at equal potential. Therefore, no current flows through that resistor or between the points A and B shown in Figure 2(a). Then the circuit of Figure 2(a) can be bisected to represent the even-mode as in Figure 2(b) for port 2. The voltage source is replaced by a short circuit because impedance calculation is required here to find the possibility of a perfect matching between the input port and the output ports. The characteristic impedance of any of the output ports \( Z_{out} \) should equal to the input impedance \( Z_{in} \) as seen from that port after taking effect of the \( \lambda_{ef}/4 \) transmission line into consideration.

\[
Z_{in} = Z_{out} = Z_{ef}/(2Z_0)
\]

(1)

It is to be noted that the \( \lambda_{ef}/4 \) transmission line represents the coupled patch of length \( l \) as shown in Figure 1. Equation (1) can be rearranged to give the required value for the even-mode impedance.

\[
Z_{in} = \sqrt{2}Z_0
\]

(2)

Concerning the odd-mode’s operation, there is a zero potential along the midplane of the circuit shown in Figure 2(a). Therefore, the circuit can be bisected as in Figure 2(c), where the midplane of the circuit is shown grounded. To achieve a perfect matching between the output ports and the input port, the characteristic impedance of any of the output ports should be equal to the input impedance \( Z_{in} \) as seen from that port. This requirement can be written as in the following equation after taking effect of the \( \lambda_{ef}/4 \) transmission line into consideration.
This equation results in the following value for the isolation resistor:

\[ R = 2Z_o \]  

(4)

Having established the principles of operation of the power divider, a simple procedure can be applied to find the design parameters of the device. The width \( w_f \) of the input and output microstrip ports is determined by assuming 50 \( \Omega \) characteristic impedance and then using the well known microstrip design equations [1]. From (2) and (4), the required values for the even-mode impedance and the isolation resistor can be found:

\[ Z_{oe} = 70.7 \Omega, \quad R = 100 \Omega \]  

(3)

Using the calculated value of the even-mode impedance, width of the slot at the ground plane \( w_s \) and diameter of the microstrip coupled patch \( D_m \) can be estimated using the quasi-static analysis presented in [9, 10]. It is to be noted that when using the design equations of [10], \( 2w_c \) is to be replaced by \( D_m \).

Concerning the odd-mode impedance of the coupled region, the odd-mode equivalent circuit [Fig. 2(c)] indicated that this impedance is terminated with a short circuit. This means that the input impedance as seen from any port is independent of the odd-mode impedance. Thus, it has no effect on the operation and can take any value. This can also be justified by the fact that the presented device is an inphase power divider, which means that if the three ports of the device are perfectly matched then the two output signals are equal and inphase. This means that the even-mode is the normal mode of operation for the described device.

Radius of the circle used to effectively open ended the slotline at the ground plane \( r_s \) can be chosen to be around twice of the microstrip width \( w_f \) according to the guidelines presented in [6].

3. RESULTS AND DISCUSSIONS

To validate the presented method, a UWB inphase power divider was designed and manufactured. Rogers RO4003C, with thickness = 0.508 mm, \( \varepsilon_r = 3.38 \), and tangent loss = 0.0023, was used as a substrate. Values of the design parameters \( (l, w_f, D_m, w_s, r_s, \text{ and } R) \) calculated using the outlined design procedure and optimized using the software HFSS are 8.5 mm, 1.18 mm, 2.3 mm, 0.7 mm, 3 mm, and 100 \( \Omega \), respectively. The overall dimension of the device is 2 cm \( \times \) 3 cm.

The device was tested via simulations (using HFSS) and measurements (using a vector network analyser). The simulated and measured S-parameters of the power divider are shown in Figure 3. It is to be noted that due to symmetry, performance of the output port 3 is exactly similar to that of port 2. Therefore, the results for port 3 are not shown in Figure 3.

The results in Figure 3 reveal that the power is equally divided between the two output ports with an insertion loss less than 0.5 dB across the band 3.1–10.6 GHz. Also the return loss for the input port and the output ports is better than 23 dB at the centre of the band and it is better than 10 dB for the whole ultra wideband. The isolation between the output ports is found to be more than 30 dB at the centre of the band and is better than 10 dB across the whole UWB. There is generally a good agreement between the simulated and measured results.

The phase performance of the developed device was also measured. It was found that the signals from the two output ports are inphase with less than 2.5° phase imbalance over the UWB as depicted in Figure 4.
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4. CONCLUSION
An ultra wideband inphase power divider has been described. The proposed device utilizes a T-microstrip junction combined with a controlled coupling between a slotted ground plane and an elliptical patch at the centre of the T-junction. An isolation resistor is connected across the slot in the ground plane to enhance the isolation between the output ports. The simulated and measured return losses at the three ports of the device and isolation between its output ports have shown an ultra wideband performance. The measured result has also shown high phase stability with less than 2.5° phase imbalance between the output signals.

ABSTRACT: A CPS-fed printed pentagonal-loop antenna with high gain, and broad beam is proposed. This antenna has not only the high gain of 11.7 dB, but also the low cross-polarization level less than −44 dB. Furthermore, the broad beam width with more than 60° HPBW (Half Power Beam Width) in the E plane is able to avoid the demand of precise main beam alignment. With these advantages, it can be used in the low power density applications of rectenna systems and radio frequency front-ends. For measuring, a balun from the CPS to the microstrip line is designed. The measured results are in agreement with the simulated ones. © 2009 Wiley Periodicals, Inc. Microwave Opt Technol Lett 51: 1188–1191, 2009; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24292

Key words: high gain; broad beam; low cross-polarization level; CPS (coplanar stripline); balun

1. INTRODUCTION
Antennas are the key components of radio frequency front-ends. High gain, low cross-polarization, broad beam and ease of integration are demanded with the rapid progress of communication and rectenna systems.

The antenna with high gain can be used in low power density applications. A variety of microstrip antenna structures are proposed to obtain higher gain, such as adding reflecting metal plane [1, 2], the double-layered structure [3]. In recent years, the electronic band gap (EBG) and photonic band gap (PBG) structure are used to enhance the gain [4]. Table 1 compares the performances of antennas above.

Generally, high gain antennas have comparably narrow beams which require a precise main-beam alignment between the transmitter and the receiving antenna. So a high gain antenna with broad beam is preferable.

Ease of integration is demanded by radio frequency front-ends. The unipolar transmission-line based on a coplanar stripline (CPS) has been developed for monolithic microwave integrated circuits (MMIC) [5]. CPS characteristics include low loss, small dispersion, compact size, low parasitic discontinuity, and ease of mounting lumped components. Its balanced structure makes CPS a good candidate for circuit design in such devices as printed dipole antenna feed [6], rectennas [7], unipolar mixers [8]. Figure 1 shows the performance of the CPS with the field distribution being drawn. w and s are the stripline width and gap between the two striplines. CPS has more design freedom. The characteristic im-

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Three-Way Parallel-Coupled Microstrip Power Divider With Ultrawideband Performance and Equal-Power Outputs

Amin M. Abbosh, Senior Member, IEEE

Abstract—A compact equal-power three-way divider with ultrawideband performance is presented. The proposed device utilizes simple, three parallel-coupled microstrip lines. In order to enable the use of practical gaps between the tightly coupled lines, slotted ground plane and lumped capacitors, which are connected symmetrically between the two sidelines and the centerline, are utilized. The conformal mapping technique is employed to find the dimensions of the device. The simulated and measured output power, return loss, and isolation show that the proposed divider operates well across the frequency band from 4 to 11 GHz.

Index Terms—Coupled transmission lines, multiway power divider (PD), three-way power divider (PD).

I. INTRODUCTION

POWER dividers (PDs) are key components in many microwave systems. They are widely used in antenna arrays, power amplifiers, phase shifters, and vector modulators, etc.

Several methods have recently been proposed to design equal-power wideband multiway dividers [1]–[6]. Those methods include a modified Wilkinson PD [1], spatial architectures [2], [3] multilayer broadside-coupled structures [4], [5], and a combination of parallel-coupled lines and isolation resistors [6].

The modified Wilkinson divider [1] requires the use of thin film technology to fabricate very narrow gaps between the utilized coupled lines. Concerning the spatial structures [2], [3], they suffer generally from a low isolation between the output ports. A fractional bandwidth of 20% was achieved by using a single-section structure [2], whereas a 100% fractional bandwidth was achieved by using three sections with Chebyshev transformer [3]. Regarding the multilayer broadside-coupled approach [4], [5], it achieves an ultrawideband (UWB) performance. However, they require the use of five conductive layers and four layers of substrates. The use of two sets of parallel-coupled structures and isolation resistors [6] requires also the use of bond wires and narrow gaps. The performance shows a limited bandwidth.

Building an equal-power three-way PD using a coupled structure requires very low odd-odd mode impedance and very high even-even mode impedance. For the conventional parallel-coupled microstrip lines, those two requirements can only be achieved when the gap between the coupled lines is extremely small. That value for the gap makes the development of the device using the printed circuit board (PCB) technology impractical. Thus, it is not a surprise when reviewing the literature not to find any PCB-based equal-power three-way divider designed using the conventional three parallel-coupled microstrip lines.

In this letter, an equal-power three-way divider based on three parallel-coupled microstrip lines is proposed. To enable the use of practical spacing between the coupled lines, two independent strategies are employed: Firstly, lumped capacitors are used symmetrically to connect the centerline with each of the sidelines. As proven previously with directional couplers [7], those capacitors increase the effective odd-odd mode capacitor, and thus, decrease the odd-odd mode impedance without any impact on the even-even mode circuit. Secondly, a slotted ground is used underneath the coupled structure to decrease the effective even-even mode capacitor, and thus to increase the even-even mode impedance with negligible impact on the odd-odd mode circuit. A design method based on the conformal mapping is used to find the physical dimensions. The simulated and measured results validate the success of the proposed method.

II. PROPOSED METHOD

The proposed three-way parallel-coupled microstrip PD is shown in Fig. 1. The input signal is applied at port 1, whereas the direct output signal appears at port 2. Since the utilized structure is of a backward coupled type, the coupled outputs appear at ports 3 and 4. The ports 5 and 6 are isolated ports, and thus they are terminated with 50 Ω resistors.

The design of a three-way PD with equal outputs requires a tight coupled structure, and thus low odd-odd mode impedance and high even-even mode impedance. To achieve the first requirement, lumped capacitors are connected between the centerline and the two sidelines, whereas a slotted ground is used...
to realize the second requirement. The utilized capacitors and slotted ground are shown in Fig. 1.

The structure of Fig. 1 can be analyzed using a quasi-static transverse electromagnetic approach. A cross-sectional view of the structure showing the equivalent capacitors per unit length is depicted in Fig. 2. For the utilized parallel-coupled structure, it is assumed that the mutual coupling between the sidelines is negligible compared with the coupling between any of the sidelines and the centerline. This assumption is based on the analysis presented in [5] and [8].

The presented device is designed to have equal output power at its three outputs. Thus, the two symmetrical microstrip sidelines (#2 and #3 in Fig. 2) have equal dimensions. Thus, all the calculated values of the mode impedances and dimensions for transmission line #2 are equal to those of transmission line #3.

For the even-even mode, the three lines are excited at the same polarity. Thus, the virtual walls shown in Fig. 2 behave as E-walls. Therefore, the even-even mode capacitors of the centerline (#1) and the two sidelines (#2 and #3) are

\[ C_{1ee} = C_{g1}, \quad C_{2ee} = C_{3ee} = C_g. \]

For the odd-odd mode, the two sidelines are excited in phase, whereas the centerline is out of phase. Hence, the virtual walls shown in Fig. 2 behave as E-walls, and, thus the odd-odd mode capacitors of the three coupled lines are

\[ C_{1oe} = C_{g1} + 4C_{ad}, \]
\[ C_{2oe} = C_{g2} + 2C_{g1}. \]

In the third mode, i.e., even-odd mode, the centreline is grounded, whereas the two sidelines are excited with opposite polarity. Under these conditions, the even-odd mode capacitor of the two sidelines can be calculated from Fig. 2 as

\[ C_{2eo} = C_{3eo} = C_g + C_{ad} + \frac{C_g}{4}. \]

The different capacitors used in (1)–(4) are shown in Fig. 2: \( C_{g1} \) is the added capacitor per unit length. \( C_{g1} \) and \( C_g \) are the effective capacitors per unit length between the centerline and sidelines, respectively, and the ground, whereas \( C_{ad} \) is the capacitor per unit length between the coupled lines.

The characteristic impedance of each of the three lines at any mode (say \( Z_{\ell} \) at the \( \ell \)-th mode) can be found from [5]

\[ Z_{\ell} = \frac{\sqrt{\varepsilon_r}}{\varepsilon_r} \]

\( \varepsilon_r \) is the speed of light in free space, \( C_{g1} \) is the \( \ell \)-th mode capacitor per unit length of the line, and \( \varepsilon_r \) is the \( \ell \)-th mode effective dielectric constant. For the utilized configuration, most of the electric field lines are within the substrate in the even-even mode. Thus, the effective dielectric constant is considered to be equal to that of the substrate \( (\varepsilon_r) \). For the odd-odd mode, the lines are distributed in the dielectric of the substrate and the free space above that substrate; hence the effective dielectric constant is taken as \( (1 + \varepsilon_r)/2 \).

It is to be noted from (2), (3) that the total odd-odd mode capacitor of each line increases due to the effect of the added capacitor \( C_{g1} \) . Thus, the odd-odd mode impedance decreases for a certain gap \( s \) as compared with the case without \( C_{g1} \). Concerning the even-even mode, the only capacitor in this mode is the one between any of the coupled lines and the ground, i.e., \( C_{g1} \) for the centerline and \( C_g \) for any of the sidelines. Hence, removing the ground underneath the coupled structure leads to a significant reduction in the even-even mode capacitors, and thus, to a significant increase in the even-even mode impedances. The removal of the ground plane underneath the coupled structure has negligible impact on the odd-odd mode capacitor due to the relatively small value of \( C_{g1} \) or \( C_g \) with respect to \( 4(C_{ad} + C_g) \) or \( 2(C_{ad} + C_g) \) which are not affected by the slot.

Using the analysis presented in [5] and [9], it is possible to show that the coupling between the two sidelines is negligible if the three-line coupled structure is designed such that

\[ Z_{oe} = \sqrt{Z_{1oe}Z_{2oe}} \]
\[ Z_{ee} = \sqrt{Z_{1ee}Z_{2ee}} \]
\[ Z_{ed} = \sqrt{Z_{1ee} + Z_{1oe}Z_{2ee} + Z_{2oe}} \]

The different capacitors used in (1)–(4) are shown in Fig. 2: \( C_{g1} \) is the added capacitor per unit length. \( C_{g2} \) and \( C_g \) are the effective capacitors per unit length between the centerline and sidelines, respectively, and the ground, whereas \( C_{ad} \) is the capacitor per unit length between the coupled lines.

The characteristic impedance of each of the three lines at any mode (say \( Z_{\ell} \) at the \( \ell \)-th mode) can be found from [5]

\[ Z_{\ell} = \frac{\sqrt{\varepsilon_r}}{\varepsilon_r} \]

\( \varepsilon_r \) is the speed of light in free space, \( \ell \)-th mode capacitor per unit length of the line, and \( \varepsilon_r \) is the \( \ell \)-th mode effective dielectric constant. For the utilized configuration, most of the electric field lines are within the substrate in the even-even mode. Thus, the effective dielectric constant is considered to be equal to that of the substrate \( (\varepsilon_r) \). For the odd-odd mode, the lines are distributed in the dielectric of the substrate and the free space above that substrate; hence the effective dielectric constant is taken as \( (1 + \varepsilon_r)/2 \).

It is to be noted from (2), (3) that the total odd-odd mode capacitor of each line increases due to the effect of the added capacitor \( C_{g1} \) . Thus, the odd-odd mode impedance decreases for a certain gap \( s \) as compared with the case without \( C_{g1} \). Concerning the even-even mode, the only capacitor in this mode is the one between any of the coupled lines and the ground, i.e., \( C_{g1} \) for the centerline and \( C_g \) for any of the sidelines. Hence, removing the ground underneath the coupled structure leads to a significant reduction in the even-even mode capacitors, and thus, to a significant increase in the even-even mode impedances. The removal of the ground plane underneath the coupled structure has negligible impact on the odd-odd mode capacitor due to the relatively small value of \( C_{g1} \) or \( C_g \) with respect to \( 4(C_{ad} + C_g) \) or \( 2(C_{ad} + C_g) \) which are not affected by the slot.

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\[ Z_{oe} = \sqrt{Z_{1oe}Z_{2oe}} \]
\[ Z_{ee} = \sqrt{Z_{1ee}Z_{2ee}} \]
\[ Z_{ed} = \sqrt{Z_{1ee} + Z_{1oe}Z_{2ee} + Z_{2oe}} \]

To complete the design procedure, the relation between the different capacitors used in (1)–(5), and the physical dimensions depicted in Fig. 1 is needed. Using the conformal mapping technique and following the procedure used in [10], one can show:

\[ C_{g1} = 2\varepsilon_r \frac{K_{01}}{K_{11}} \]
\[ C_g = 2\varepsilon_r \frac{K_{02}}{K_{12}} \]
\[ C_{ad} = \varepsilon_r (1 + (1 + \varepsilon_r) \frac{K_{03}}{K_{13}}) \]

\[ K_1 = \frac{1 + \exp \left[ \frac{-\pi u_{1}}{2\varepsilon_r} \right]}{1 + \exp \left[ -\pi u_{1} \right]} \]
\[ K_2 = \frac{1 + \exp \left[ \frac{-\pi u_{2}}{2\varepsilon_r} \right]}{1 + \exp \left[ -\pi u_{2} \right]} \]
that is available from Johanson Technology (USA) has the value of 0.1 pF. Thus, one capacitor of 0.1 pF is connected at the center between each of the sidelines and the centerline of the manufactured device. To compensate for the slight difference between the calculated and available capacitor and to get the best possible performance, values of the design parameters were optimized using the software CST Microwave Studio. The final values are: \( s = 0.11 \text{ mm}, \quad \nu = 1.75 \text{ mm}, \quad \nu = 1.55 \text{ mm}, \quad \nu = 8 \text{ mm}, \quad l = 3.7 \text{ mm}, \quad \nu = l = 0.1 \text{ pF}, \) which are reasonably close to the initial calculated values. The developed device (Fig. 3) has an overall dimension of 2.5 cm \( \times \) 2.5 cm.

The performance of the designed device according to the simulations and measurements is shown in Fig. 4. The power at each of the output ports is 4.77 dB \( \pm 1 \) dB across the band from 4 to 11 GHz in the simulations, and from 4.3 to 11 GHz in the measured results. Fig. 4 also shows the return loss at any of the four ports (due to symmetry \( \phi_{11} = \phi_{22} \) and \( \phi_{33} = \phi_{44} \)), and the isolation between the output ports (due to symmetry \( \phi_{32} = \phi_{43} \)). The return loss at any of the four ports is better than 16 dB, whereas the isolation between the output ports is better than 12 dB across the whole investigated band from 3 to 11 GHz.

The simulated and measured performances agree well with each other as shown in Fig. 4. There is a slight difference between them, especially at the upper band. This difference can be attributed to the variation in the utilized lumped capacitors, which are not expected to have a constant value across the whole investigated band.

IV. CONCLUSION

A compact equal-power three-way PD has been presented. The proposed divider utilizes a simple three parallel-coupled microstrip lines. To enable the use of practical dimensions, slotted ground underneath the coupled structure, and lumped capacitors connecting the coupled lines are utilized. The simulated and measured results of the device prove its UWB performance.

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The noise of the locked ILFD at 1 MHz offset is

Figure 6 shows the measured phase noises of the injection-refer-

cences are found at each

Figure 4 shows the operation frequency and locking range

dependence on input power for the ILFD biased at \( V_{dd} = 0.7 \) V,

while varying \( V_{tune} \) from 0 to 0.9 V. The locking range increases

with injection power. It has a useful high band, and an external

injected signal power of 0 dBm provides an operation range of

3.9–6.1 GHz. At \( V_{tune} = 0 \) V, the locking range is 2.1 GHz,

from 3.9 to 6.0 GHz. It has an excited low band at \( V_{tune} = 0.9 \)

V; this is caused by a dual-resonance ILO.

Figure 5 shows the measured output spectra of the high-band

ILFD before and after the locked conditions. The locked output

spectra show a lower phase noise than that of the free-running

ILFD. The output power at 9.4 GHz of ILFD is \(-7.764\) dBm.

Figure 6 shows the measured phase noises of the injection-refer-

ence and high-band ILFD. Although the input signal is with a

power of 0 dBm, the phase noise of input signal at 1 MHz offset

is \(-126.73\) dBc/Hz, after external power injection, the phase

noise of the locked ILFD at 1 MHz offset is \(-122.49\) dBc/Hz.

Figure 7 shows the operation frequency and locking range

dependence on input power for the ILFD biased at \( V_{dd} = 0.7 \) V,

while varying \( V_{tune} \) at 1.2 and 2 V. Two locking range charac-


teristics are found at each \( V_{tune} \). An external injected signal

power of 0 dBm provides a low-band operation range of 1.7–2

GHz. At \( V_{tune} = 1.2 \) V, the locking range is 0.3 GHz, from 1.7
to 2 GHz. An external injected signal power of 0 dBm provides a

high-band operation range from 5.1 to 5.6 GHz. Figure 8 shows

the high-band harmonic suppression versus injection input

power. For the input level of 0 dBm, the ILFD shows a func-

tional suppression of \(-30.66\) dBm and third harmonic suppress-

sion of \(-38.12\) dBm.

4. CONCLUSIONS

This letter proposes a fully integrated dual-band differential

input/output CMOS LC-tank injection-locked frequency doubler

that uses a cross-coupled VCO with a dual-resonance first-har-

monic ILO and frequency doubler with differential outputs. To

our knowledge, this is the first reported dual-band ILFD with
dual-resonance LC tank. The ILFD consumes low power and has

high/low operation range from the incident frequency 3.9/1.7
to 6.1/2 GHz to provide output signals with the frequency from

7.8/3.4 to 12/2.4 GHz, while tuning the oscillation fre-

quency. The ILFD can also provide dual-band locking range at

a fixed high-tuning bias. The measured result shows that the

prototype ILFD is useful for RF circuit application.

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ULTRA-WIDEBAND THREE-WAY POWER DIVIDER USING BROADSIDE-COUPLED MICROSTRIP-COPLANAR WAVEGUIDE

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ABSTRACT: A compact three-way power divider with ultra-wideband (UWB) performance is presented. The proposed device utilizes a simple and low-cost broadside-coupled microstrip-coplanar waveguide structure. The conformal mapping technique is used to find the dimensions of the device. The simulated and measured output power, return loss, and isolation validate the suitability of the proposed divider for the UWB applications that operate across the frequency band from 3.1 GHz to more than 10.6 GHz. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 54:196–199, 2012; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.26492

Key words: power divider; planar power divider; three-way power
divider; broadside-coupled structure

1. INTRODUCTION

Power dividers are key components in many microwave sys-
tems. They are widely used in mixers, antenna arrays, power
amplifiers, phase shifters, and vector modulators, to name a few.
The most well-known three-way power divider is the Wilkin-
son divider [1]. However, a three-way Wilkinson divider faces a
serious packaging problem as it requires a three-dimensional
floating common node. To overcome that problem, different
configurations have been proposed [2, 3]. However, the
measured performance of the proposed configurations shows a limited bandwidth.

The multilayer broadside-coupled microstrip-slot-microstrip approach has recently been proposed for the development of wideband three-way power dividers [4, 5]. It uses five conductive layers interleaved with four layers of substrates. In this article, a simple configuration that uses only two layers of substrates is proposed to design a compact three-way power divider with ultra-wideband (UWB) performance. A broadside-coupled microstrip-coplanar waveguide (CPW) structure is utilized to achieve the required performance. The conformal mapping technique is used to calculate the required dimensions of the device. The simulated and measured results of an equal-power three-way prototype divider show an output power of $-4.77 \pm 1$ (nominal value = $-4.77$ dB) for each of the three output ports across the band from 3.1 to 11.5 GHz. The proposed divider exhibits better than 16 dB return loss at its ports with more than 13 dB isolation between the three output ports across the ultra-wide frequency band from 3.1 to 10.6 GHz.

2. CONFIGURATION AND DESIGN

The configuration of the proposed three-way power divider is shown in Figure 1. It consists of three conductive layers interleaved with two substrates. The input and one of the output ports are located at the middle layer of the structure [Fig. 1(b)], whereas the other two output ports in the form of CPW are at the top [Fig. 1(a)] and bottom [Fig. 1(c)] layers. The other two output ports at the top and bottom coupled layers labeled as matched ports in Figures 1(a) and 1(c) have no power output, and thus, they are terminated with 50 $\Omega$ surface mount resistors to absorb any reflected signal from the three output ports labeled with #2, 3, and 4 in Figure 1(d). To have one common ground for the whole structure, the ground plane of the top and bottom layers indicated in Figures 1(a) and 1(c) is connected with each other by soldering them with the ground terminal of subminiature A connectors at the four ports of the device.

Assume that it is required to have a power ratio of $abc$ at the output ports 2, 3, and 4, respectively. The required coupling factors (CF1 and CF2) between the microstrip patch at the middle layer, which is connected to the input port, and the CPW at the top and bottom layers should be [5]:

$$CF_1 = \sqrt{b/(a+b+c)}$$

$$CF_2 = \sqrt{c/(a+b+c)}.$$

If equal-power output is required at the three output ports, that is, the power from each of the three output ports is $1/3$ of the input power, and thus, a power ratio of $1:1:1$ is required, $CF_1 = CF_2 = CF = \sqrt{1/3}$, which is equivalent to $-4.77$ dB.

The proposed structure shown in Figure 1 is asymmetrical. Thus, the $c$- and $\pi$-mode used to find the characteristics of the coupled structure. With $CF = -4.77$ dB, it is possible to show that the $c$- and $\pi$-mode impedances of the different coupled layers are 96.5 $\Omega$ and 25.9 $\Omega$, respectively [4]. The dimensions of the coupled structure offering the required $c$- and $\pi$-mode impedances can be determined by using the conformal mapping technique for the asymmetrical quasi-static model of broadside-coupled microstrip/CPW.

The excitations needed to generate the two modes and distribution of the electric field lines between the coupled layers are shown in Figure 2. For the $c$-mode, the three layers are excited in-phase, whereas in the $\pi$-mode, the top and bottom layers are out-of-phase with respect to the middle layer.

Assuming a quasi transverse electromagnetic propagation, the $c$- and $\pi$-mode impedances of the coupled lines can be determined from the effective capacitances per unit length of the lines and the phase velocity on the lines [6]. Those capacitances are shown for the two fundamental modes in Figure 3. The $c$-mode capacitances for the microstrip ($C_{mc}$) at the middle layer and the CPW ($C_{cpw}$) at the top and bottom layers are equal to
The \( p \)-mode capacitances for the microstrip (\( C_{mp} \)) and the CPW (\( C_{cp} \)) are equal to

\[
C_{mp} = 2C_{mg} + 4C_{mc}; C_{cp} = C_{cg} + 2C_{mc}.
\] (2)

Using the conformal mapping technique, the capacitances shown in Figure 3 are calculated as a function of dimensions of the coupled structure [7]

\[
C_{mg} = 2\varepsilon_0\varepsilon_r K(k_1)K(k_1)
\] (3)

\[
C_{cg} = 2\varepsilon_0\left(\frac{\varepsilon_r}{\varepsilon_0}\right)^1 K(k_2)K(k_2) + 2\varepsilon K(k_3)K(k_3)
\] (4)

\[
C_{mc} = \varepsilon_0\varepsilon_r w_m + w_c
\] (5)

where \( K(k) \) and \( K'(k) \) are the first kind elliptical integral and its complementary, respectively, \( w_m, w_c, \) and \( w_s \) are the widths of the coupled structures as shown in Figure 1, and \( h \) is the thickness of the substrate.

The characteristic impedances of the coupled CPW at the top and bottom layers and the microstrip at the middle layer at the \( c \)-and \( p \)-modes have the following relation with the mode capacitances [6].

\[
Z_{ij} = \sqrt{\varepsilon_r (c_0 C_{ij})},
\] (9)

where the subscript \( i \) refers to the line (m for microstrip and c for CPW) and \( j \) refers to the mode (c for \( c \)-mode and \( p \) for \( p \)-mode), \( c_0 \) is the velocity of light in free space, and \( \varepsilon_r \) is the dielectric constant of the substrate.

As the utilized structure shown in Figure 1 is asymmetrical (the middle layer is a microstrip line, whereas the top and bottom layers are CPW), the CF between the middle layer and any of the other two layers is [5]

\[
CF = \frac{Z_{cc} - Z_{cc}}{\sqrt{(Z_{mc} + Z_{mc})(Z_{cc} + Z_{cc})}}
\] (10)

In deriving (10), the three layers are assumed to have characteristic impedance equal to \( Z_0 = \sqrt{Z_{mc}Z_{mc}} = \sqrt{Z_{cc}Z_{cc}} = 50 \) \( \Omega \).
4. CONCLUSIONS

A compact three-way power divider with UWB performance has been presented. The proposed divider utilizes a simple broadside-coupled microstrip/CPW structure. The simulated and measured results of the developed device have shown equal three-way power division, good return loss, and isolation across the band from 3.1 to 10.6 GHz.

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THE EXPERIMENTAL STUDY OF MUTUAL COUPLING IN CONFORMAL ARRAY

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ABSTRACT: The problem of the analyzing mutual coupling for the conformal array in practice is addressed. Experiment is designed for the sectoral cylinder conformal array composed of 12 cavity-backed stacked microstrip antennas. Based on the measurement element complex patterns, the phase pattern calibration technique and least squares method are adopted to calculate the mutual coupling matrix (MCM) for the conformal array. It is shown that the actual array pattern modified by MCM agrees well with the ideal desired pattern, which confirms the validity and accuracy of the calculation of the MCM. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 54:199–203, 2012. View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.26491

Key words: conformal array; microstrip antenna; mutual coupling; pattern synthesis

1. INTRODUCTION

In recent years, conformal phased array systems have developed considerably because of their capability of complying with the requirements for aerodynamic shape, low antenna radar cross section, wide scan range, etc. [1, 2]. Microstrip patch antennas are often used as the array element because of their thin profile, light weight, and low cost [3]. Using the stacked patches could overcome the narrowband drawbacks of microstrip antenna [4, 5]. Therefore, the stacked microstrip antenna is extensively investigated to design the conformal array element.

In published literatures, several methods have been applied for analysis of conformal array; however, those methods assume...
Figure 8 shows the measured radiation patterns including the copolarization and cross-polarization in the H-plane (x–z plane) and E-plane (x–y plane). It can be seen that the radiation patterns in x–z plane are nearly omnidirectional for the two frequencies.

4. CONCLUSIONS
In this article, a novel multiresonances printed monopole antenna by using a pair of Γ-shaped arms and an H-shaped slot in the bottom of microstrip feedline on the ground plane, much wider impedance bandwidth can be produced, especially at the higher band.

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adjusting circuit. The device is manufactured showing a wide bandwidth in terms of return loss, isolation, power division, and a differential phase shift of 90° across the frequency band of 3–8 GHz. Its compact size and good performance makes it suitable for use in wideband-balanced amplifiers. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 54:300–305, 2012; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.26519

**Key words:** power divider; phase shifter; wireless via; microstrip circuits

1. **INTRODUCTION**

Power dividers are essential components of many microwave subsystems [1]. They are used in RF front ends of microwave transceivers (in amplifiers and mixers) as well as in beamforming networks of array antennas. One of the most prominent examples of two-way power divider is the Wilkinson power divider [1]. Its importance in microwave subsystems comes from its properties such as high return losses of its three ports and a high isolation of its two output ports. These properties make the Wilkinson divider also useful as a power combiner [1]. Its functions of power division and combining can be utilized in developing high power solid-state microwave sources [2].

In its basic configuration, the Wilkinson divider offers an in-phase signal division. However, there are applications, such as balanced amplifiers, where the two amplifiers require the power division and combination with a 90° phase difference [1]. In practice, this function can be delivered using two 3 dB quadrature couplers with one of their ports terminated in a matched load [1]. The role of the two couplers is that they improve the return loss performance of the balanced amplifier as well as deliver graceful degradation, in case one of the amplifiers fails. Typically, the couplers are developed in planar (microstrip, stripline, or coplanar waveguide) technology. However, the matched load is realized as a coaxial termination to achieve wideband performance. This leads to a cumbersome and expensive process of developing a balanced amplifier.

A solution to this problem is described in Refs. 3–5 where 3 dB quadrature couplers are replaced by a quadrature power divider (QPD). In the presented solution, QPD is formed by the Wilkinson divider accompanied by additional circuits, so that the two output ports offer a differential phase shift of 90°. The usual challenge, as in the previous solution with the 3 dB quadrature couplers, is to achieve the differential phase shift of 90° between the output ports over a wide operational bandwidth. In this case, one has to overcome the fundamental behavior of low-pass transmission lines that, for a given length, provide an increasing negative phase shift with frequency. As shown in Refs. 3–5, this behavior can be counter counted by high-pass type transmission lines that, for a given length, provide an increased positive phase shift with frequency. If the two lines offer the phase shift change with frequency at the same pace, then the differential phase shift introduced by the two lines can be made constant. The line with the positive phase shift utilizing lumped inserts in an ordinary microstrip line is called the metamaterial (MM) transmission line.

Using the approach outlined in Refs. 3–5, obtaining a differential phase shift of 90° can be practically realized by various combinations of negative and positive phase shifts of the lines (i.e., 90° = 60°(−30°), 90° = 45°(−45°)), with preference given to the combination that leads to the smallest deviation from 90° over a specified frequency band.

One practical inconvenience of high-pass (MM) microwave transmission lines is that adding lumped inductances and capacitances to an ordinary microstrip line leads to the bulky hybrid approach of realizing such structures.

An alternative approach to obtaining a stable differential phase shift is via the phase reversal. Examples of this approach are transmission lines utilizing wired or wireless vias [6]. Their advantageous feature is that they offer a wideband performance [7]. The shortcoming of the wireless vias, such as those presented in Ref. 6, is that they require the use of multilayer structures when realized in microstrip-slot technology.

In this article, we use a double wireless via which is designed on a single substrate with two conductor coated sides to form the Wilkinson type QPD. The utilized via exhibits an ultrawideband performance and makes the Wilkinson type QPD fully planar and compatible with ordinary microstrip circuits.

2. **DESIGN**

The configuration of the proposed QPD is shown in Figure 1. It consists of a Wilkinson power divider, conventional microstrip transmission lines, and a double wireless via as a phase adjusting circuit. The Wilkinson power divider is chosen in the proposed QPD, as it can equally separate the input power into two output ports and also provide good isolation between them. A conventional microstrip transmission line and a double wireless via acting as a phase adjusting circuit are connected to two output ports of the divider to achieve broadband quadrature phase difference. The proposed design is similar to Refs. 3 and

![Figure 1](https://example.com/image1.png)

**Figure 1** Configuration of (a) the broadband quadrature power divider employing and (b) a double wireless via
Figure 2  Single-stage Wilkinson power divider (a) structure and (b) simulated s-parameter in dB. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 3  Double-stage Wilkinson power divider (a) configuration, (b) CST layout, and (c) simulated s-parameter in dB. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
However, in this article, a phase adjusting circuit is formed by a double wireless via instead of metamaterial transmission lines in Refs. 3 and 5. Because the new configuration does not include any chip inductors or capacitors, as required in Refs. 3 and 5, it is fully planar.

The design of the proposed QPD is divided into two parts, the Wilkinson power divider and a double wireless via, which are then combined to form the device. The design of the Wilkinson divider follows well-established rules. Here, we investigate single- and double-stage varieties to be used in QPD.

The design aims at an ultrawideband performance across the band from 3.1 to 10.6 GHz. In the design, Rogers RT6010 substrate with dielectric constant of 10.2, tangent loss of 0.0023, and thickness of 0.635 mm is used. The design is aided with the commercially available full-wave EM analysis and design tool CST Microwave Studio.

At first, a single-stage Wilkinson power divider is designed following the circuit configuration as shown in Figure 1(a). The single-stage design requires a 100-Ω resistor for the isolation of the output ports. The simulated results of the one-stage power divider indicates a return loss and isolation greater than 10 dB and equal power division (3 dB ± 0.7) within the band from 3 to 10 GHz.

Because of use of a substrate with high dielectric constant, the overall dimensions of the divider are very small. These small dimensions form a motivation for designing a double-stage power divider for the proposed QPD structure.

The configuration of a double-stage power divider and its simulated results are shown in Figure 3. By comparing the results in Figures 2 and 3, it is apparent that for return loss greater than 15 dB and equal power division (3 dB ± 0.7) within the band from 3 to 9 GHz.

The factor $\beta_{ef}$ is the effective phase constant. By following the equations described in Ref. 8, the initial dimensions of the structure shown in Figure 1(b), $D_m$, $D_c$, $D_s$, $l_1$, $l_2$, and $l_3$ can be obtained. Further adjustments can be made using CST Microwave Studio.

The following equation can be used to work out dimensions for a constant 90° differential phase shift [7]:

$$\Delta\Phi = 180' - 4 \arctan \left( \frac{\sin(\beta_{ef} l_3)}{1 - \beta_{ef}^2 \cos(\beta_{ef} l_3)} \right) + \beta_{ef} l_m = \theta$$

The configuration of a double-stage power divider and its simulated results are shown in Figure 3. By comparing the results in Figures 2 and 3, it is apparent that for return loss greater than 15 dB, the frequency range is from 4 to 9.5 GHz for the double stage, whereas it is limited to the band from 5 to 9 GHz for the single stage. The isolation is also improved to 20 dB for most of the band. Because of satisfactory performance both of them can be considered for inclusion in QPD. However, in this article, the experimental results are only reported for the QPD with the double-stage Wilkinson divider.

The design of a double wireless via acting as a phase adjusting circuit of QPD follows the ideas described in Ref. 8. It uses only a single substrate, and, thus, it forms a uniplanar structure with its input and output ports located on the same side of the substrate.

As shown in Figure 1(b), each section of the wireless via is composed of a microstrip patch on the top layer, and a coplanar waveguide structure is implemented on the bottom layer. The level of electromagnetic coupling between the top and the bottom layers is controlled by adjusting the dimensions of the slot in the bottom layer. The two broadside coupled sections are connected together through a microstrip transmission line on the bottom layer. The property of this double wireless via is that it offers a constant phase difference over a wide frequency band when compared with a suitably chosen length $l_m$ of an ordinary microstripline. The differential phase shift can be controlled by the coupling factor, $CF$ between the top layer and the bottom layer for both sections [7]. This can be accomplished by adjusting the minor axis of the elliptical patches, $D_m$ and $D_c$, and slot, $l_s$, which have values close to quarter of the effective wavelength at the center of the passband.

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$$\theta$$

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Figure 5  The photograph of manufactured QPD. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
Here, the double wireless via for 90° differential phase shift is designed assuming Rogers RT6010 substrate with dielectric constant of 10.2, tangent loss of 0.0023, and thickness of 0.635 mm. The aim is to achieve the differential phase shift of 90° across UWB of 3.1–10.6 GHz. After optimization using the software, the dimensions in mm are Dm = 3.5, Dc = 1.5, Ds = 5.8, l1 = 4.3, l2 = 4.1, and l3 = 4.9, and l∞ = 18.6. The simulated structure excluding the reference line is very compact with dimensions of $10 \times 14 \text{ mm}^2$. The simulated performance of the double wireless via acting as the QPD phase adjusting circuit is presented in Figure 4.

As observed in Figure 4, for the simulated performance of the designed via, the 10 dB return loss bandwidth is from 3 to 10.3 GHz. A return loss of not less than 20 dB is achieved from 3.5 to 9.3 GHz. The device features very low insertion loss which is less than 0.5 dB within 3–10.2 GHz. The phase imbalance of the 90° differential phase shift is $\pm 5°$ across the same band when compared with a 50-$\Omega$ microstrip transmission line of length, $l_∞ = 18.6 \text{ mm}$. These results show that the structure is suitable to be combined with the earlier designed Wilkinson power divider to obtain the quadrature phase response of QPD over a wide bandwidth.

3. IMPLEMENTATION AND RESULTS

The broadband QPD is designed and manufactured as shown in Figure 5.

As stated earlier, here only results for the QPD using the double Wilkinson divider are reported. The prototype is fabricated using Protomat C100/HF Micro Milling Machine. For the substrate RT6010, the fabricated QPD is of dimensions $28 \times 24 \text{ mm}^2$ excluding the connectors and thus represents a compact design in the intended frequency band. The required resistor values are rounded to 91 and 240 $\Omega$. It is verified through simulations that the performance of the Wilkinson divider is not affected by the small deviations in the resistor values. The 91-$\Omega$ resistor is realized using chip resistor 0805 (1.6 $\times$ 0.8 mm$^2$), whereas the 240-$\Omega$ resistor is realized using chip resistor 0805 (2.0 $\times$ 1.5 mm$^2$). The experimental testing is investigated over the frequency band from 3 to 11 GHz. Figure 6 shows the simulated and measured results of the manufactured QPD.

The return loss is better than 10 dB for both simulated and measured results across the band from 3 to 9.8 GHz. The curve shape of the measured return loss is similar to the simulated one. However, it is slightly shifted to the low frequency above 6 GHz. The observed discrepancies between the simulated and measured results in the output power and return loss can be explained by the use of coaxial to microstrip transitions in the measurement system which were not included in the simulations. The performance of the connectors used in this case is degraded at the upper end of the investigated frequency band. The other reasons for discrepancies can be explained by manufacturing tolerances. The simulated and measured isolation is greater than 15 dB across the whole band. As shown in Figure 6(b), a good agreement between the simulated and measured results can be noticed in the differential phase shift of the QPD from 3 to 8 GHz. In this frequency range, the device achieves the differential phase shift of 90° $\pm 5°$. This has to be considered as a very impressive performance of QPD in comparison with the results reported in Refs. 3–5.

4. CONCLUSIONS

A compact quadrature power divider in uniplanar microstrip technology has been designed, fabricated, and tested. The proposed device uses the conventional Wilkinson power divider with one of its output arms equipped with a double wireless via acting as a phase adjusting circuit and the other one reserved for a reference microstrip line. According to the simulated and measured results, the prototype has a wide bandwidth in terms of return loss, isolation, power division, and a differential phase shift of 90°. Its compact size and good performance makes it suitable for use in wideband-balanced amplifiers.

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1. INTRODUCTION

The occupied size of the internal laptop computer or tablet computer applications for the penta-band wireless wide area network (WWAN) or eight-band WWAN/long term evolution (LTE) operation is generally not easy to be reduced owing to the large lower-band bandwidth required (for instance, 824–960 MHz for the GSM850/900 operation or 704–960 MHz for the LTE700/ GSM850/900 operation) [1–12]. The requirement of large lower-band bandwidth is especially more challenging for the internal antenna in the tablet computer than in the mobile handset. This is because the size of the system ground plane is generally much larger in the tablet computer than in the mobile handset [13–16], thus making the chassis mode of the system ground plane generally cannot be excited to assist in achieving a wider lower band for the embedded internal antenna.

To overcome the aforementioned problem for the internal WWAN/LTE antenna embedded in the tablet computer, we present a bandwidth-enhancement technique by using a distributed parallel resonant circuit embedded in the antenna, which does not increase the antenna’s occupied size and can lead to a dual-resonance excitation of the antenna’s lower band for the LTE700/ GSM850/900 (704–960 MHz) operation with a small size of 12 × 60 × 3.8 mm³. The distributed parallel resonant circuit is obtained by embedding a long narrow strip with its front terminal connected to one part of the antenna’s main radiator and its end section gap-coupled to the other part of the main radiator. The former provides an additional inductance and the latter contributes an equivalent capacitance, and both can lead to the generation of a parallel resonance seen at the antenna’s feeding point [17]. This parallel resonance can result in the generation of an additional resonance at the high-frequency tail of the original single-resonance mode in the antenna’s lower band, such that the original single-resonance mode can become a dual-resonance mode to provide a much wider lower-band bandwidth for the antenna. Note that as the embedded parallel resonant circuit does not increase the occupied size of the antenna, bandwidth enhancement is obtained without a sacrifice in increasing the antenna size.

Further, the antenna can provide a wide upper band for the GSM1800/1900/UMTS/LTE2300/2500 (1710–2690 MHz) operation. That is, the proposed antenna can cover the eight-band WWAN/LTE operation with an occupied size about the smallest among the related reported antenna for tablet computer or laptop computer applications [11, 14–16]. Details of the proposed antenna and the operating principle of the embedded parallel resonant circuit are described in the article. The antenna was fabricated and tested. Results of the fabricated antenna are presented and discussed.

2. PROPOSED ANTENNA

Figure 1 shows the geometry of the proposed WWAN/LTE tablet computer antenna with an embedded parallel resonant circuit. A photo of the fabricated antenna is shown in Figure 2. The antenna is formed by two portions. The first portion is printed on a 0.8-mm thick FR4 substrate of relative permittivity 4.4, loss tangent 0.02, and size 12 × 60 mm², and the second portion is a metal strip of width 3 mm and length 60 mm (t in the figure) cut from a 0.2-mm thick copper plate. The second portion is oriented perpendicular to and connected with the first portion as shown in the figure. In this study, the antenna is to be applied in a 9.7-inch tablet computer, which is currently commercially available. The antenna is mounted along the edge of the top shielding metal wall (5 × 150 mm²) of the display ground (150 × 200 mm²), which is used to support a 9.7-inch display. Note that the antenna is placed close to one corner of the shielding metal wall such that other possible internal antennas (for instance, the WLAN antenna, the multiple-input multiple-output antenna system [18], etc.) can also be mounted along the
Analysis and Design of Ultra-Wideband Unequal-Split Wilkinson Power Divider Using Tapered Lines Transformers

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Abstract An ultra-wideband unequal-split Wilkinson power divider with a 2:1 split ratio is presented. To achieve the ultra-wideband characteristics, the conventional quarter-wave arms of the divider are replaced by tapered lines. Moreover, two extra tapered transformers are incorporated at the output ports for matching purposes as the designed divider is of an unequal-split type. To obtain good isolation between the output ports, five isolation resistors are used, the values of which are determined using the simple odd-mode analysis of the Wilkinson power divider. For verification purposes, an ultra-wideband Wilkinson power divider that operates over a frequency range extending from 2 to 12 GHz is designed, simulated, fabricated, and measured. The results of the full-wave simulation and measurements verify the validity of the design procedure.

Keywords power divider, Wilkinson power divider, tapered lines, ultra-wideband

1. Introduction

There is an increased interest in the design of ultra-wideband (UWB) microwave components since the Federal Communication Commission’s approval to use the frequency range of 3.1 to 10.6 GHz for UWB applications, such as short-range indoor data transmission, microwave imaging, and through-the-wall radars. As a consequence, many researchers have recently been attracted to this field, and different microwave devices that support the use of the approved UWB frequency band have been presented. The microwave power divider is one of the key microwave components used in different wireless applications. They are extensively used in antenna feed networks, balanced mixers, and phase shifters. Thus, designing power dividers to be used in UWB systems is of utmost importance.

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The Wilkinson power divider (WPD) is one of the dividers that gained a notable significance and interest in the literature related to the design of UWB dividers. For many applications, such as phase-array systems and beam-forming networks, power dividers with unequal power division ratios are required. Many configurations to build an unequal-split WPD have been investigated (Moradian & Oraizi, 2008; Li & Wang, 2011; Wu et al., 2008, 2009a; Zhu et al., 2010, Li et al., 2009a). In Moradian and Oraizi (2008), the application of the grooved substrate was presented for the design of a 4:1 unequal-split WPD. The grooves were applied along one of the divider strips that require high characteristic impedance for the purpose of overcoming its conventional narrow width. A stub-loaded transmission line was used in Li and Wang (2011) to design WPDs with arbitrary power division ratios. A WPD operating at a frequency and its first harmonic with an unequal power dividing ratio was proposed in Wu et al. (2008). To obtain the unequal power property, four groups of 1/6 wavelength transmission lines with different characteristic impedances were needed to match all ports. In Zhu et al. (2010), a shunt-stub WPD with a uniform impedance line was introduced. Compared with the conventional divider, the output distribution ratio was controlled by the length of the shunt stubs. A 10:1 unequal-split WPD was designed in Li et al. (2009a) by replacing the high-impedance line with two coupled lines terminated by two shorts. In Wu et al. (2009a), a WPD operating at an arbitrary dual-band with an unequal power dividing ratio was presented. The asymmetric structure that consists of seven sections of transmission lines with different characteristics impedances was given to achieve the unequal power division and matching characteristics. Furthermore, to obtain an acceptable isolation, a series resistor-inductor-capacitor structure was incorporated in the proposed design.

In Oh et al. (2007) and Ko et al. (2003), wideband WPDs with unequal power divisions were built. However, the design requires the use of extremely narrow lines and a slotted ground to achieve the required high impedance for one of the branches of the WPD. The use of lumped elements was investigated in Mizuno et al. (2008); however, the achieved fractional bandwidth is only 40%, and thus, it does not suit the UWB applications. Multisection stubs were utilized with the conventional WPD to realize an arbitrary power ratio but across a narrow band (Wu et al., 2009b). In another approach, offset double-sided parallel-strip lines were used to build the Wilkinson divider with an unequal power ratio (Chen & Xue, 2007). However, the utilized structure is difficult to integrate with the other microstrip-based devices. The use of different types of shunt stubs or coupled lines results in a narrowband performance (Li et al., 2009b, 2010; Wu et al., 2010; Ahn et al., 2009).

In this article, an UWB unequal-split WPD with a 2:1 split ratio is presented. The proposed device is aimed at covering the bands from 2 GHz to 12 GHz. Tapered-line transformers are incorporated in the proposed design. The design of those transformers is achieved using the even-mode analysis of the divider. To achieve acceptable output ports matching and isolation conditions, multiple resistors are mounted between the two tapered arms of the WPD. An optimization process is carried out to find the values of those uniformly distributed resistors considering the odd-mode analysis. It should be emphasized that the present article differs from Chiang and Chung (2010) in two aspects. First, an unequal-split WPD is considered here, while an equal-split WPD was investigated in Chiang and Chung (2010), and as a consequence, two different (asymmetric) microstrip tapered lines are considered in the WPD design. Furthermore, two extra tapered transformers are designed and incorporated at the power divider’s output ports for matching purposes. Second, in this study, the values of the shunt resistors are obtained through an independent optimization process using the odd-mode equivalent
circuit of the WPD, while in Chiang and Chung (2010), the built-in optimization tool in the full-wave simulator was used to find the resistors’ values.

2. Design of UWB 2:1 WPD

An outline of the proposed device is shown in Figure 1. Each branch of the conventional divider is replaced by a single microstrip tapered-line section. Since such tapered sections have almost constant input impedance across an extremely wide bandwidth, they are used to achieve the UWB operation of the WPD. In Section 2.1 (even-mode analysis), the design of the tapered lines is presented; while in Section 2.2 (odd-mode analysis), the values of the isolation resistors are derived.

2.1. Even-Mode Analysis

The even-mode equivalent circuits for the upper and lower branches of the proposed UWB divider are shown in Figure 2.

For an unequal-split WPD, $Z_{s1}$, $Z_{l1}$, $Z_{s2}$, and $Z_{l2}$ can be found using the following equations (Pozar, 2005):

\[
Z_{s1} = Z_0 \left( 1 + \frac{1}{k^2} \right),
\]

\[
Z_{l1} = \frac{Z_0}{k},
\]

\[
Z_{s2} = Z_0 (1 + k^2),
\]

\[
Z_{l2} = kZ_0,
\]

where $k = \sqrt{P_2/P_3}$, and $P_2$ and $P_3$ are the output powers from ports 2 and 3, respectively.

Figure 2. Even-mode equivalent circuits for the UWB unequal-split WPD: (a) upper branch and (b) lower branch.
Hence, to obtain a 2:1 split ratio (i.e., \( k = \sqrt{2} \)), and using a characteristic impedance \( Z_0 \) of 50 \( \Omega \), the values of \( Z_{s_1}, Z_{l_1}, Z_{s_2}, \) and \( Z_{l_2} \) should be 75 \( \Omega \), 35.35 \( \Omega \), 150 \( \Omega \), and 70.71 \( \Omega \), respectively.

According to Chiang and Chung (2010) and Hecken (1972), the maximum input return loss (in dB) for a given tapered line used to match source impedance \( Z_s \) to load impedance \( Z_l \) is characterized by the following equation:

\[
|RL_{\text{input}}|_{\text{max}} = -20 \log \left[ \tanh \left( \frac{B}{\sinh B} \right) \ln \left( \frac{Z_l}{Z_s} \right) \right], \tag{2}
\]

where \( B \) is a predefined design parameter used to determine the tapered line curve. It should be mentioned here that larger values of \( B \) result in lower reflection at the input port. However, increasing \( B \) will demand a wider tapered line width and longer length.

After choosing the value of \( B \) in order to achieve a desired input return loss, the exponential tapered line characteristic impedance is calculated using the following equation (Chiang & Chung, 2010; Hecken, 1972):

\[
\ln \left( \frac{Z(z)}{Z_s} \right) = 0.5 \ln \left( \frac{Z_l}{Z_s} \right) \left[ 1 + G(B, 2 \left( \frac{z}{d} - 0.5 \right)) \right], \tag{3a}
\]

where

\[
G(B, \xi) = \frac{B}{\sinh B} \int_{0}^{\xi} I_0 \left( B \sqrt{1 - \xi'^2} \right) d\xi'. \tag{3b}
\]

\( Z(z) \) in Eq. (3a) represents the characteristic impedance of the tapered line at point \( z \), and \( I_0(x) \) represents the modified zero-order Bessel function. The tapered line length \( d \) is a predefined variable chosen appropriately to achieve the desired maximum input return loss.

As noted above, the conventional quarter-wave transformers are replaced by their equivalent tapered-line transformers, considering \( (Z_{s_1}, Z_{l_1}) \) and \( (Z_{s_2}, Z_{l_2}) \), in order to achieve the UWB characteristics. Moreover, two extra tapered line transformers are designed to match the output ports to 50 \( \Omega \). The source and load impedances that are considered in the design of these matching tapered transformers are \( (Z_{l_1}, Z_0) \) and \( (Z_{l_2}, Z_0) \).

### 2.2. Odd-Mode Analysis

The odd-mode analysis is carried out to obtain the isolation resistors’ values needed to achieve the optimum output ports isolation and output ports matching conditions. Figure 3 shows the equivalent odd-mode circuit of the proposed divider (Qaroot et al., 2010).

First, each tapered line transformer will be subdivided into \( K \) uniform electrically short segments with length \( \Delta z = d/K \). The \( ABCD \) matrix for each section of the tapered line (considering the upper branch) shown in Figure 3 is calculated as follows (Pozar, 2005):

\[
\begin{bmatrix}
  A & B \\
  C & D \\
\end{bmatrix} = \begin{bmatrix}
  A_1 & B_1 \\
  C_1 & D_1 \\
\end{bmatrix} \begin{bmatrix}
  A_2 & B_2 \\
  C_2 & D_2 \\
\end{bmatrix} \cdots \begin{bmatrix}
  A_i & B_i \\
  C_i & D_i \\
\end{bmatrix}, \tag{4}
\]
where the $ABCD$ parameters of the $i$th segment are (Pozar, 2005)

$$A_i = D_i = \cos(\Delta \theta), \quad (5a)$$

$$B_i = Z^2((i - 0.5)\Delta z)C_i = jZ((i - 0.5)\Delta z) \sin(\Delta \theta), \quad (5b)$$

$$\Delta \theta = \frac{2\pi}{\lambda} \Delta z = \frac{2\pi}{c} f \sqrt{\varepsilon_{\text{eff}}} \Delta z. \quad (5c)$$

The effective dielectric constant $\varepsilon_{\text{eff}}$ of each section is calculated using the well-known microstrip line formula in Pozar (2005). Then, the total $ABCD$ matrix for the upper branch can be calculated as follows (Pozar, 2005):

$$[ABCD]_{\text{Total}} = [ABCD]_{R'_N} \cdot [ABCD]_{\text{first section}} \cdots$$

$$\cdot [ABCD]_{(N-1)\text{th section}} [ABCD]_{R'_1} \cdot [ABCD]_{\text{Nth section}}. \quad (6)$$

It is worth mentioning here that the isolation resistors are distributed uniformly (a resistor every $d/N$ distance, where $d$ is the tapered line length, and $N$ is the number of used resistors). Finally, and as illustrated in Figure 3, the following equation can be written:

$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{\text{Total}} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}. \quad (7)$$

Setting $V_2 = 0$ leads to the following equation:

$$Z_{\text{in}}^o = \frac{V_1}{I_1} = \frac{B}{D}. \quad (8)$$
For perfect output port matching, the following condition should be satisfied:

\[
\Gamma_{\text{out}}(f_j) = \frac{Z_{\text{in}}^o(f_j) - Z_{l_1}}{Z_{\text{in}}^o(f_j) + Z_{l_1}},
\]

(9)

where \( f_j \) denotes the frequencies at which Eq. (9) is calculated. Here, a frequency increment of 1 GHz is used within the frequency range 2 to 12 GHz. So, for perfect output ports matching over the design frequency range, the following error function is considered (Shamaileh et al., 2011):

\[
\text{Error}_{\text{out}} = \sum_{j=1}^{M} |\Gamma_{\text{out}}(f_j)|^2.
\]

(10)

This optimization problem is solved using “fminunc.m” MATLAB routine (The Math-Works, Natick, Massachusetts, USA), where \( R'_1, \ldots, R'_N \) are the optimization variables to be determined. It should be noted here that the same procedure carried out in Eqs. (4) through (10) is repeated in order to obtain the optimum resistors’ values \( R'_1, \ldots, R'_N \), which minimize the output reflection for the lower branch in Figure 3. Finally, the overall resistance values can be calculated as follows (Qaroot et al., 2010):

\[
[R_1 \ R_2 \ \cdots \ R_N] = [R'_1 \ R'_2 \ \cdots \ R'_N] + [R''_1 \ R''_2 \ \cdots \ R''_N].
\]

(11)

3. Simulation and Experimental Results

Figure 4 represents the layout of the proposed UWB 2:1 WPD (without the isolation resistors). Considering a Roger RT5870 substrate (Rogers Corporation) with a relative permittivity of 2.33, a thickness of 0.508 mm, and a loss tangent of 0.0012, the lengths of each tapered WPD arm and output ports matching transformers needed to achieve an acceptable input/output ports matching conditions are set to 28 mm and 27 mm, respectively. Those lengths are approximately equal to the lengths of the conventional uniform quarter-wave transformers at 2 GHz. Besides, the design parameter \( B \) was set to 5.5, which corresponds to a maximum input return loss of 56.55 dB.

Figure 5 shows the effect of uniformly distributing three resistors (a resistor every 9.3 mm), four resistors (a resistor every 7 mm), and five resistors (a resistor every 5.6 mm) on the input/output ports matching parameters (\( S_{11}, S_{22}, \) and \( S_{33} \)), as well as the isolation between the two output ports (\( S_{23} \)). It should be pointed out here that the simulation results were obtained using the method of moments (MoM) based full-wave simulator.

![Figure 4. Layout of the 2:1 UWB WPD without the isolation resistors (dimensions in mm).](image)
Isolation and input/output ports matching parameters of the designed UWB WPD with different numbers of isolation resistors. (color figure available online)

IE3D (Mentor Graphics PCB Design Software, 2006). Moreover, the isolation resistors’ values in the three scenarios are obtained following the optimization procedure mentioned in Section 2.2. Table 1 shows the resulting resistors in each case.

It is clearly seen in Table 1 that the resistors’ values increase in magnitude in all cases but that of three resistors. This reflects the divergence of the optimization engine toward minimizing the error expressed in Eq. (10) when only three resistors are used.
The error achieved when using three resistors was more than 0.9, which is considered very high as an output reflection coefficient, whereas the error is around 0.1 when using four and five resistors. The divergence in the three-resistor case was translated into the resistors’ variation listed in Table 1. Table 2 represents the error obtained in each case along with the obtained error value in both branches.

As shown in Figure 5, when using three resistors, an acceptable return loss (below 10 dB) at the input port and port 2 is achieved across the entire frequency range of interest. However, poor isolation and output return loss at port 3 are obtained at some frequency bands. On the other hand, better performance is clearly seen in the case of using four and five isolation resistors.

Figure 6 shows the simulated transmission parameters $S_{21}$ and $S_{31}$. Lower transmission loss is obtained in the case of using four and five resistors because of the optimal response achieved by the optimization engine in those two cases. In both scenarios, $S_{21}$ equals $-1.76$ dB ($\pm 0.8$ dB), while $S_{31}$ equals $-4.77$ dB ($\pm 1$ dB) over the frequency range of 2 to 12 GHz.

For verification purposes, the UWB unequal-split WPD with five isolation resistors is implemented over the same substrate previously mentioned. The practical surface-mount device (SMD) resistor values that are used in the fabrication are $R_1 = 100 \, \Omega$, $R_2 = 200 \, \Omega$, $R_3 = 330 \, \Omega$, $R_4 = 470 \, \Omega$, and $R_5 = 510 \, \Omega$. Figure 7 shows a photograph of the fabricated divider, while Figure 8 shows the simulated and measured scattering parameters.

As shown in Figure 8(a), both simulated and measured results show an acceptable input port matching (below $-10$ dB) over the frequency range of 2 to 12 GHz. Furthermore, the output ports matching parameters $S_{22}$ and $S_{33}$ are also below $-10$ dB over the same frequency range. Moreover, the simulations and the measurements results illustrated in Figure 8(b) show an acceptable isolation (below $-10$ dB) over the design frequency range. The simulation results for the transmission parameters are close to their theoretical values; $S_{21}$ is $-1.77$ dB ($\pm 0.5$ dB) in the frequency range of 2 to 12 GHz, and $S_{31}$ is

<table>
<thead>
<tr>
<th>No. of resistors</th>
<th>$R_1$ ((\Omega))</th>
<th>$R_2$ ((\Omega))</th>
<th>$R_3$ ((\Omega))</th>
<th>$R_4$ ((\Omega))</th>
<th>$R_5$ ((\Omega))</th>
</tr>
</thead>
<tbody>
<tr>
<td>Three resistors</td>
<td>22.34</td>
<td>651.41</td>
<td>428.765</td>
<td>—</td>
<td>—</td>
</tr>
<tr>
<td>Four resistors</td>
<td>91.57</td>
<td>210.73</td>
<td>360.52</td>
<td>476.74</td>
<td>—</td>
</tr>
<tr>
<td>Five resistors</td>
<td>97.75</td>
<td>203.41</td>
<td>328.87</td>
<td>477</td>
<td>515.8</td>
</tr>
</tbody>
</table>

Table 1

Optimized values of isolation resistors used in simulations of the 2:1 UWB WPD

The error values in the optimization as calculated from Eq. (10) for upper and lower branches of the device

<table>
<thead>
<tr>
<th>No. of resistors</th>
<th>Upper branch</th>
<th>Lower branch</th>
</tr>
</thead>
<tbody>
<tr>
<td>Three resistors</td>
<td>0.92</td>
<td>0.97</td>
</tr>
<tr>
<td>Four resistors</td>
<td>0.1</td>
<td>0.14</td>
</tr>
<tr>
<td>Five resistors</td>
<td>0.093</td>
<td>0.091</td>
</tr>
</tbody>
</table>
Figure 6. Transmission parameters $S_{21}$ and $S_{31}$ for the 2:1 UWB WPD with different numbers of isolation resistors. (color figure available online)

$-4.77 \text{ dB (±1 dB)}$ over the same frequency range. The measured transmission parameters show acceptable characteristics except for a slight increase of the insertion losses at the upper end of the investigated band. That increase, as well as the discrepancies between the simulated and measured results is thought to be due to the connectors, the tolerance in the values of the five resistors, conductor and dielectric losses, and radiation losses.

Figure 9 shows the phase imbalance of the designed divider, which clearly illustrates an in-phase performance with less than $2^\circ$ imbalance over the entire frequency range that extends from 2 to 12 GHz. It is to be noted here that for applications that require compact
Figure 8. Simulated and measured scattering parameters for the fabricated UWB 2:1 WPD. (color figure available online)
design, meandered structures, for example, can be utilized to decrease the overall length of the proposed divider.

4. Conclusions

The design of an UWB unequal-split WPD using tapered lines has been presented. The design of the UWB tapered lines is obtained from the even-mode analysis of the WPD, whereas the isolation resistors are calculated through an optimization process using the odd-mode equivalent circuit. Three scenarios are presented in such a way that the effect of using three, four, and five isolation resistors on enhancing the isolation between the two output ports, as well as achieving optimum output ports matching over the frequency range of 2 to 12 GHz, is studied. For verification purposes, an UWB unequal-split WPD, with a 2:1 split ratio, and five isolation resistors, is fabricated and measured. The good agreement between both simulation and measurement results and the design target proves the validity of the design procedure.

References


Three-way signal divider with tunable ratio for adaptive transmitting antenna arrays

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Abstract: A tunable wideband three-way signal divider that is useful in adaptive transmitting arrays is presented. The design of the proposed device is based on using three stepped-impedance coupled microstrip lines that have controllable coupling factors, and thus a variable signal ratio at the three output ports. The variation in the coupling factors is achieved by using two varactor diodes that are connected between the central coupled line and each of the side lines. The biasing voltages of the varactor diodes are used to control their capacitors and thus to achieve the required output signal ratios. A signal-flow analysis is used to predict the performance of the proposed device, whereas the conformal mapping technique is used to obtain the required dimensions of the coupled structure and the varactors’ capacitors. The validation of the proposed design method is tested by simulations and measurements. The results indicate the possibility of changing the signal ratio across a wide range of values with more than 10 dB return loss at the four input/output ports and isolation between the output ports across more than 85% fractional bandwidth.

1 Introduction

Adaptive arrays are widely used for many purposes, such as to achieve extremely low-sidelobes or to optimise the radiation pattern for certain required characteristics concerning, for example, the gain or signal-to-noise ratio subject to the existence of noise sources or external interferences [1–4].

Two of the important applications of transmitting arrays that need pattern optimisation are the multiple-input multiple-output (MIMO) systems and retrodirective arrays [1–3]. In MIMO systems, the pattern, for example, is synthesised to transmit data to a single user while placing nulls in the directions of other users [2, 3]. In retrodirective arrays, the transmitting patterns are adapted to optimise the transmission subject to some received signal or noise distribution [1]. They find applications in many commercial and military systems, such as electronic traffic and toll management, wireless power transmission, advanced digital mobile communication systems, where limited tracking without a priori knowledge of the source is required, and many more [2, 4].

The adaptation of the array’s pattern can be realised by real-time phase and/or amplitude weighting of the signals fed to the elements of the array. In this paper, a device is proposed to enable changing the amplitude of the signals in an efficient and simple manner.

The conventional scenario to implement amplitude–weight pattern optimisation on a transmitting array is to use as many variable gain amplifiers as transmitting antenna elements (Fig. 1a). The gain of those amplifiers is varied as per the amplitude weights required to generate a certain pattern. Obviously, this solution requires the use of a large number of variable-gain amplifiers. The other proposed solution depicted in Fig. 1b is to have one fixed-gain amplifier for the whole system, whereas the signal is distributed among the antenna elements using a multiway signal divider with tunable amplitude ratio at its output ports. This configuration is useful for low-power levels. For high-transmitting-power levels, the configuration shown in Fig. 1c that includes an easy-to-design fixed-gain amplifier for each antenna element and a variable-ratio multiway signal divider to control the amplitude weights is more realistic.

Reviewing the literature shows that there are several methods proposed to build tunable two-way signal dividers and couplers [5–8]. Most of the available designs of tunable two-way dividers are based on the theory proposed in 1958 [5], which states that the series connection of two 3-dB directional couplers using a variable phase shifter results in a tunable power divider. Concerning multiway (three-way or more) signal dividers, several configurations have been proposed to build wideband and cost-effective devices [9–18]. However, all those devices have fixed signal division, and thus, are not suitable as an implementation tool for the pattern optimisation of adaptive arrays based on amplitude weights.

In [15, 19], novel couplers and power dividers utilising coupled structures are proposed. In those designs, a chip capacitor is connected between the coupled lines at the centre of the structure to achieve a certain value for the coupling factor without the need to use narrow gaps or lines that are difficult to manufacture. If that capacitor is replaced with a varactor diode, it could be possible to change the coupling factor of the structure, and thus the amplitude ratio of the output signals, by changing the biasing voltage applied to the diode. Thus, signal dividers
with tunable amplitude ratio can be realised using that approach with a properly designed coupled structure.

In this paper, a wideband three-way signal divider with planar structure and variable amplitude ratio at its three output ports is proposed. The device is based on three parallel-coupled microstrip lines. Unlike the available tunable two-way dividers that need cascading multiple couplers or dividers and variable phase shifters, the tenability of the proposed three-way divider is achieved by the direct control of the coupling factor between those three lines. That target is realised by connecting two varactor diodes between the coupled lines. A complete design method based on the signal flow diagrams and quasi-static approach is presented. The designed device has a compact size, planar structure and tunable amplitude ratio across more than 85% fractional bandwidth.

2 Theory

Assume that there are three stepped-impedance coupled lines as shown in Fig. 2. The coupled stepped-impedance structure is proven to be useful in extending the bandwidth of coupled devices [19, 20]. Those lines can be arranged in a parallel-coupled structure for uniplanar configuration [15] or broadside-coupled structure for multilayer configuration [13, 16]. Each of the three coupled lines includes three sections. The two side sections are similar with a length of \( l_1 \), whereas the central section has a length \( l_2 \) as depicted in Fig. 2. Assume that based on the dimensions of the side sections, the electromagnetic coupling factor between the three lines forming them is equal to \( c_s \). The central section is assumed to have a coupling factor \( c_{12} \) with the upper-side line and \( c_{13} \) with the lower-side line.

The structure shown in Fig. 2 can be analysed using a combination of even–odd mode analysis using the quasi-static approach and signal-flow diagrams [19, 20]. In the following analysis of the three-line coupled structure, it is assumed that all the six ports of the structure are perfectly matched. Moreover, the coupling between the side lines is assumed to be negligible. The input signal at port 1 is divided between the three output ports 2, 3 and 4 and no signal appears at the ports (5 and 6), which can then be terminated by matched loads. The division of the signal between the three output ports follows the following scattering parameters, which are derived using the signal flow of multiport devices at a certain frequency with effective wavelength \( \lambda_e \).

\[
S_{21} = \frac{\alpha_1 + \beta_2 \beta_6}{1 - \alpha_1 \alpha_4 - \alpha_1 \alpha_6} + \alpha_3
\]
\[
S_{31} = \frac{\alpha_1 + \beta_2 \beta_6}{1 - \alpha_1 \alpha_4 - \alpha_1 \alpha_6} + \alpha_3
\]
\[
S_{41} = \frac{\beta_1 \beta_2 \beta_6}{1 - \alpha_1 \alpha_4 - \alpha_1 \alpha_6}
\]

where

\[
\alpha_1 = \frac{j c_s \sin(2\pi l_1/\lambda_e)}{\sqrt{1 - 2 c_s^2 \cos(2\pi l_1/\lambda_e)} + j \sin(2\pi l_1/\lambda_e)}
\]
\[
\alpha_2 = \frac{j c_s \sin(2\pi l_1/\lambda_e)}{\sqrt{1 - c_s^2 \cos(2\pi l_1/\lambda_e)} + j \sin(2\pi l_1/\lambda_e)}
\]
The performance of the proposed three-way signal divider is calculated using the derived model for different values of the lengths and coupling factors. It is found that the following equation derived previously in [19, 20] concerning the relation between the length of the side sections \(l_s\) and the length of the central section \(l_c\) for a wideband performance of four-port devices is still valid in the proposed six-port device for a wideband operation

\[ l_c = 2l_s \]  

As with the analysis in [20], the total length of the coupled structure is equal to one-third of the effective wavelength at the centre of the required operational bandwidth.

The calculated performance for different values of the coupling factors is shown in Figs. 3–5 across one-octave frequency band. It is clear from Figs. 3 and 4 that fixing the value of \(c_s\) and \(c_{21}\) (or \(c_{31}\)) while changing the value of \(c_{31}\) (or \(c_{21}\)) causes a significant variation in the level of signals emerging from the output ports 4 and 3 (or 2), whereas the level of the signal at port 2 (or 3) varies slightly. Those results can be explained by the fact that increasing the coupling factor from the central line to any side line increases the signal emerging from the output port connected to that side line, and consequently decreases the remaining signal emerging from the output port connected to the central line (port 4). To be able to change the signal level at any ratio between the three output ports, the coupling factors \(c_{21}\) and \(c_{31}\) are to be changed together as indicated in the results of Fig. 5.
The signal ratio can also be controlled by changing the coupling factors of the first and third sections, while fixing that factor for the central section. However, there is a practical implication of that approach in that the number of varactor diodes is larger and this means an increase in the losses because of the parasitic elements of those diodes in addition to complicating the biasing circuit.

### 3 Design

The tunable ratio of signals can be achieved as explained earlier by changing the coupling factors of the central coupled section. The ideal scenario for that change to happen while keeping the perfect matching between the whole coupled structure and the impedance ($Z_o$) of the input/output ports, is to vary both of the odd–odd and even–even mode impedances of the central section ($Z_{oo}$ and $Z_{ee}$) so that at any coupling factors, and thus at any signal ratio, the following equation holds [20]

$$\sqrt{Z_{ee}Z_{oo}} = Z_o = 50 \, \Omega$$

Practically, this target can be implemented by using five varactor diodes to control the coupling factors of the central section. Two of those diodes are connected from the central line to the two side lines, whereas the three remaining diodes are connected between each of the three-coupled lines and the ground. By properly changing the biasing voltages of the five varactor diodes, the perfect matching can be maintained across the required range of signal ratios. However, this solution complicates the design as it requires five different biasing voltages to control the capacitance of the utilised varactor diodes.

The other solution is to use only two diodes connecting the central line to the two side lines. Those diodes are used to vary the odd–odd mode impedance of the central section. The even–even mode impedance is designed to have a reasonable high value that can guarantee a proper matching. To avoid using narrow gaps or lines that makes the manufacturing process difficult, the high value for the even–even mode impedance can be realised by using a slotted ground underneath the central coupled structure [20]. The dimensions of the coupled structure are calculated so that a perfect matching is achieved at three-equal signal outputs, that is, signal ratio of 1:1:1. For other ratios, the capacitor of each of the varactor diodes can take values that are above or below the required value at the 1:1:1 ratio. A compromise between the performances at the different signal ratios can be obtained using a suitable full-wave electromagnetic simulator so that an acceptable performance can be maintained at all the possible signal ratios.

The final adopted design is depicted in Fig. 6a. The top layer contains three parallel-coupled microstrip lines, whereas a slotted ground plane is located at the bottom layer directly underneath the central coupled section. In order to feed the diodes, a proper biasing circuit is included in the design as revealed in Fig. 6a.

The dimensions of the coupled structure of Fig. 6a can be calculated using the even–even, odd–odd and even–even mode approach [15, 20, 21]. Assuming a quasi-transverse electromagnetic propagation, the electrical characteristics of the coupled lines (mode impedances and thus the coupling factors) can be calculated from the effective per unit length capacitances of lines and the phase velocity on the lines. In turn, the values of different mode capacitors can be calculated using conformal mapping techniques as explained in [21, 22]. The main parameters that define the values of those capacitors are the thickness and dielectric constant of the substrate ($h$ and $\varepsilon_r$), and the dimensions of the coupled structure ($s, w, w_s$ and $w_c$) depicted in Fig. 6a. Based on that approach, the required dimensions of the three sections and the required varactors’ capacitors are found assuming a signal ratio of 1:1:1. The final dimensions are then obtained via the optimisation feature of the software HFSS. Assuming the use of the substrate Rogers RT6010 ($\varepsilon_r = 10.2, \ h = 0.635 \, \text{mm}$), the optimised dimensions in (mm) for a signal divider operating across the band from 2 to 5 GHz are $l_1 = 2.7, \ s = 0.21, \ l_2 = 4.8, \ w_s = 4.9, \ w = 0.89, \ w_c = 0.44$ and varactors’ capacitors at 1:1:1 signal ratio $C_{v1} = C_{v2} = 1 \, \text{pF}$. The overall dimensions of the device are 30 mm $\times$ 30 mm.

The range across which the signal level at each of the output ports can be controlled depends on the range of the capacitor of the utilised varactor diodes. A large capacitor range means wide range of signal levels that can be achieved at each output port. To control the signal at any of the output ports by a ratio (maximum value/minimum value) of 3, the required capacitors ($C_{v1}$ and $C_{v2}$) of the utilised varactor diodes should cover the range from 0.4 to 1.8 pF. Those values are evenly distributed around the required value for a signal ratio of 1:1:1 and this justifies the use of the 1:1:1 ratio as the starting design point.

### 4 Results and discussions

The performance of the designed device is tested via full-wave electromagnetic simulations. Also, a prototype is
developed (Fig. 5b) and tested. To realise the optimised range of values for the capacitors of the varactor diodes, the hyper-abrupt junction diode MA46H201 is used in the developed device. The maximum biasing voltage for this diode is 20 V.

The simulated performance of the designed three-way signal divider is shown in Figs. 7–9 for varactor capacitors that are changed from 0.4 to 1.8 pF. It is clear from the results of Fig. 7, which includes limited steps of the capacitor values for clarity of the figure, that the ratio of the signals at the three output ports can be controlled across a wide range. The deviation in the values is around ±0.5 dB from the median value of each ratio across the band from 2 to 5 GHz. It is interesting to see that the general variations in the simulated signal levels at the three output ports (Fig. 7) agree well with the predictions using the derived theoretical model (Fig. 5) despite the fact that the derived model assumes an ideal substrate, perfect matching at all the ports and no coupling between the side lines.

The simulated reflection coefficient at all the ports of the device (owing to symmetry $S_{44} = S_{11}$) in all the cases is below $-10$ dB as shown in Fig. 8. There is also a good isolation, which is more than 13 dB between the output ports 2 and 4 or ports 3 and 4 in all the investigated cases as depicted in Fig. 9. The isolation between the output ports is an important factor to prevent any signal reflected from any of the antenna elements of the array depicted in Fig. 1b from affecting the performance of the other antennas in the array. The isolation between the output ports 2 and 3 is more than 10 dB in most of the cases. The only case which causes the isolation to be around 9 dB is when the biasing voltages at the varactor diodes are close to zero and thus the varactors have the largest capacitor value.

The measured performance of the developed device is shown in Figs. 10–12 for biasing voltages of the two utilised diodes that change from 2.5 to 15 V. The results of Fig. 10 concerning the ratio of the signals at the three output ports, which are shown for small number of biasing voltages steps for clarity of the figure, agree well with the simulated results (Fig. 7). The selected biasing voltages shown in Fig. 10 are chosen based on the technical data of the diodes so that their capacitors are roughly equal to those used to obtain the simulated results of Fig. 7 to enable a comparison between them. The obtained results indicate the possibility of controlling the signal ratio at the three output ports.

**Fig. 7** Simulated signal level at the three output ports for different values of the varactors’ capacitors

**Fig. 8** Simulated reflection coefficient at the input and output ports for different values of the varactors’ capacitors

**Fig. 9** Simulated isolation between the output ports for different values of the varactors’ capacitors

**Fig. 10** Measured signal level at the three output ports for different values of the varactors’ biasing voltages
ports by a factor of around 2.8, which is close to the value 3 used in the design procedure. The deviation in the output signal values is around ±1 dB from the median value of each ratio across the band from 2 to 5 GHz.

The measured reflection coefficient at the input port (port 1) of the device is below −10 dB as shown in Fig. 11. For the other two output ports, the reflection coefficient is less than −8 dB. For the configuration of transmitting array depicted in Fig. 1b, the reflection coefficient of the input port is the most important parameter among the reflection coefficients of the four ports as it quantifies the level of signal reflected back from the three-way signal divider to the amplifier. To avoid any disturbance to the operation of the amplifier, that signal should be as low as possible, which occurs when the input reflection coefficient is very low. For the output ports (2, 3 and 4), there is no signal entering the output ports when the device is used with transmitting antenna array depicted in Fig. 1b. Thus, the reflection coefficients at those ports are not critical parameter in the design compared with the input reflection coefficient.

As indicated in Fig. 12, there is also a good isolation of more than 12 dB between the output ports 2 and 4 or ports 3 and 4 in all the investigated cases. The isolation between the output ports 2 and 3 is more than 10 dB in most of the cases. As in the simulations, the only case that causes the isolation to be around 9 dB is when the biasing voltages at the varactor diodes are close to zero causing the varactor’s capacitors to be at their maximum values. In this case, the coupling factor between the two side lines, which is assumed to be zero in the design, has a certain non-zero value that causes a slight degradation in the isolation between the two-side lines.

Comparing the simulated results in Figs. 7–9 with the measured results in Figs. 10–12 reveals a good agreement between them. The slight differences are thought to be owing to the parasitic elements of the utilised varactor diodes and the imperfect isolation between the biasing circuit and the coupled structure.

The other important factor to evaluate the proposed device is the phase performance. For the application intended in the work, that is, the amplitude–weight pattern optimisation of a transmitting array, the difference in the phase between the three output ports of the signal divider should have a certain fixed value across the band of interest so that no additional variable phase compensation techniques are needed. Concerning the designed device, the differential phase of the ports 2 and 3 with respect to port 4 are calculated using the simulation tool and measured. Snapshots of the results are shown in Fig. 13 for three cases that represent the two ends of the operating conditions of the device and the central design working condition. It is clear from the results that the ports 2 and 3 have an average 90° in the simulations and 92° in the measurements fixed phase difference with respect to port 4. The deviation in that phase difference is around ±6° in the simulations and ±8° in the measured results across the investigated band. This result can be explained by the fact the coupled structures between the two output ports (2 and 3) and the input port (1) are in the form of a quadrature coupler, whereas the output port 4 is directly connected with the input port. Thus, the two ports (2 and 3) are in phase, whereas a fixed 90° phase shift is expected between them and the other output port (4).
5 Conclusion

A wideband three-way signal divider with tunable amplitude ratio has been presented. The proposed device utilises a planar parallel-coupled configuration and two varactor diodes. Through changing the biasing voltage of the varactor diodes, the coupling factor is changed and thus the signal level at the three output ports is varied. The signal flow of multiport devices is used to explain the performance of the device, whereas the conformal mapping is used to find its dimensions. The simulated and measured results have shown the possibility of achieving wide range of signal levels at the three output ports across more than 85% fractional bandwidth. The proposed device is a good candidate for pattern optimisation of low-power transmitting antenna arrays.

6 References

Design of wideband six-port network formed by in-phase and quadrature Wilkinson dividers

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Abstract: The design of a novel wideband six-port network constituted by in-phase and quadrature Wilkinson power dividers is presented. To achieve wideband operation of the quadrature divider, the device uses a 90° phase shifter in the form of a double vertical wireless interconnect that utilises microstrip to coplanar waveguide transitions. The performance of the designed in-phase and quadrature dividers and the entire six-port network is assessed via full-wave electromagnetic simulations and measurements. The obtained results show that the designed six-port network offers good performance in terms of amplitude and phase characteristics across the frequency band of 3.5–9 GHz.

1 Introduction

Six-port networks are widely used in many fields, such as reflectometers in microwave measurement systems [1–7], direct phase shift keying modulators of a transmitter [8–10] and as correlating demodulators in direct conversion receivers [8–13]. Recently, there has been a considerable interest in integrated six-port networks with wide operational bandwidth. Using a traditional approach, a six-port for the use in measurement or communication subsystems can be constructed with three couplers and one power divider [5–7]. The design of an integrated six port using the conventional configuration with wideband performance was presented in [5]. The design was accomplished in a multi-layer microstrip-slot technology in which two dielectric substrates with a common ground plane were used. The device was not only fully integrated but was also very compact in size. It provided a very good operation across an ultrawide frequency band of 3.1–10.6 GHz. However, these positive attributes came at a price. The design required a special effort of devising a new type of a power divider which could be compatible with the multi-layer couplers. In practice, the multi-layer approach creates a challenge to manufacturing of couplers and dividers, as the performance in multi-layer structures is very sensitive to any misalignments or air gaps [14]. Advanced manufacturing tools are required to align various layers without any air gaps to achieve best electrical performance.

Some of the challenges in designing six-port networks formed by couplers are addressed in [8]. The question was posed whether a traditional six port can be designed using other components than the couplers. It was shown that an equivalent performance to that of the traditional six port (formed by three couplers and one power divider) can be achieved using entirely power dividers but with some of them equipped with 90° phase shifters. An extra motivation for such a new configuration was to overcome the difficulty of designing couplers at millimetre-wave frequencies. They demonstrated their idea by designing a wideband six-port network operating at microwave frequencies. The concept of using dividers to design the six port was fully proved. However, the presented implementation was incomplete, as in the presented solution, the required 90° phase shifters were realised by delay lines referenced against sections of microstrip lines. The delay lines could only deliver a narrowband phase shift. In this paper, we provide the missing solution for wideband 90° phase shifters and describe the full design of an ultrawideband six-port network formed entirely by in-phase and quadrature Wilkinson dividers. We show that the required 90° phase shifters can be obtained using a double wireless vertical interconnect utilising a microstrip to coplanar waveguide (CPW) transition, which is described here. This design has the advantage of eliminating one terminating load, which helps in reducing some errors. Moreover, the design can be implemented on a single-layer board. It is to be noted that the multi-layer technology is used to achieve the wideband performance in [5], whereas the simple printed circuit board (PCB) technology is utilised in the proposed design.

2 Design procedure

Fig. 1 shows the block diagram of the conventional six-port network followed by its emulation proposed in [8]. The conventional device is formed by three couplers, denoted by the letter Q and one in-phase power divider, denoted by the letter D. As seen in Fig. 1a, the conventional design has an additional port that needs to be terminated with a 50 Ω load. The non-conventional six-port presented in Fig. 1b is taken from [8]. It is redrawn in this way so as to show that
it is formed by in-phase and quadrature Wilkinson dividers. In the two six-port depicted in Fig. 1, the ports are numbered this way so that an equivalence between the two can be observed in terms of the distributed signals when ports 1 and 2 are fed by complex signals $a$ and $b$.

Fig. 1 offers an ample analysis of operation of these devices when port 1 is fed with a complex signal $a$ and port 2 is fed by a complex signal $b$. The complex signals emerging from ports 3–6 are denoted as $S_3$, $S_4$, $S_5$ and $S_6$. They are linear combinations of the input signals $a$ and $b$. Those $S$-parameters are obtained assuming an ideal operation of couplers, dividers and phase shifters in Figs. 1a and b. In these figures, the symbol $j$ stands for $\sqrt{-1}$ and represents the 90° phase shift of a complex microwave signal in the phasor notation. For the conventional six-port (Fig. 1a), the required $S$-parameters can be rewritten as

$$S_3 = \frac{-ja}{2} \left[ \frac{b}{a} + 1 \right]$$  \hspace{1cm} (1a)

$$S_4 = \frac{-ja}{2} \left[ \frac{b}{a} - j \right]$$  \hspace{1cm} (1b)

$$S_5 = \frac{a}{2} \left[ \frac{b}{a} - j \right]$$  \hspace{1cm} (1c)

$$S_6 = \frac{a}{2} \left[ \frac{b}{a} - 1 \right]$$  \hspace{1cm} (1d)

whereas for the non-conventional six port (Fig. 1b) they are given as

$$S_3 = \frac{-ja}{2\sqrt{2}} \left[ \frac{b}{a} + j \right]$$  \hspace{1cm} (2a)

$$S_4 = \frac{-ja}{2\sqrt{2}} \left[ \frac{b}{a} + 1 \right]$$  \hspace{1cm} (2b)

$$S_5 = \frac{a}{2\sqrt{2}} \left[ \frac{b}{a} - j \right]$$  \hspace{1cm} (2c)

![Fig. 1](image1.png)

**Fig. 1**  Block diagram of a six-port network as proposed in [8]

a Conventional

b Non-conventional

![Fig. 2](image2.png)

**Fig. 2**  Double-stage Wilkinson power divider

a Configuration

b Fabricated
By comparing the two sets of equations, it is evident that the expressions in the square brackets in (1a)–(1d) and (2a)–(2d) are identical. The differences include the magnitudes. The ones appearing in (2) are smaller by the factor of $\sqrt{2}$ than the ones in (1). This difference occurs because in the non-conventional six-port configuration, the signals undergo a 3 dB power loss. This power loss occurs during the combining process in the Wilkinson dividers connected to ports 3–6. The remaining difference is in the sign in expressions for the signal $S_6$. Otherwise, the equivalence in operation of the two devices is apparent.

Now the task is to obtain practical realisation of the device of Fig. 1b. As observed, in Fig. 1b this device includes six in-phase Wilkinson dividers and four quadrature Wilkinson dividers. The design of Wilkinson dividers is already well established in the microwave literature [15, 16]. For example, double-stage dividers can easily be designed in microstrip technology using the guidelines given in [15, 16]. The challenge is to obtain the design of a wideband quadrature Wilkinson divider. This device is formed by a conventional Wilkinson divider with a 90° phase shifter in its output arms. The design of a wideband 90° phase shifter that complements the in-phase divider to achieve the quadrature operation is presented here. The proposed design is based on the so-called double vertical interconnect utilising microstrip to CPW transitions.

In the undertaken approach, each component (power divider and phase shifter) are designed independently using CST Microwave Studio v2010. Then they are integrated into a six-port network. The design assumes the use of a double-sided Rogers RT6010 PCB, with a relative dielectric constant of 10.2 and a loss tangent of 0.0023, 0.635 mm thickness and 17 $\mu$m conductive coating.

### 2.1 Double-stage Wilkinson power divider

Fig. 2 shows a double-stage Wilkinson power divider for use in the wideband six-port network of Fig. 1b.

The double-stage Wilkinson power divider is designed at the centre frequency of 6 GHz for operation in the 3–10 GHz band. It is known that a better performance can be achieved with more stages, but the overall dimension of the power divider would become excessively large. As a substrate with high dielectric constant is assumed in this design, the overall dimensions of the double-stage Wilkinson divider are quite small and is 15 × 20 mm excluding the connectors. The prototype of Wilkinson power divider is fabricated using Protomat C100/HF Micro Milling Machine. The required resistor values shown in Fig. 2a are rounded to 91 and 240 Ω. The 91 Ω resistor is realised using chip resistor 0603 (1.6 × 0.8 mm), whereas the 240 Ω resistor is realised using chip resistor 0805 (2.0 × 1.5 mm). The experimental testing is investigated over the frequency band from 3 to 11 GHz. Fig. 3 shows the simulated and measured results of the manufactured Wilkinson divider.

As shown in Fig. 3, a good agreement between the simulated and measured results can be noticed in the power division of the Wilkinson divider from 3 to 9 GHz. The return loss is better than 10 dB for both simulated and measured results across the band from 3 to 9.5 GHz. The simulated and measured isolation is greater than 15 dB across the whole band.
2.2 Quadrature divider

The detailed layout of the quadrature Wilkinson divider is shown in Fig. 4. The device is formed by the combination of the earlier design of double in-phase Wilkinson divider (Fig. 2) and wideband 90° phase shifter. The design of the 90° phase shifter is adopted from [17]. The device uses a two section of broadside-coupled microstrip/CPW as indicated in Fig. 4a. The design uses only a single substrate with its input and output ports located on the same side of the substrate. As shown in Fig. 4, each section of the phase shifter is composed of a microstrip patch on the top layer, and a CPW structure on the bottom layer. The two broadside coupled sections are connected together through a short section of microstrip transmission line on the bottom layer. The property of this circuit is that it offers a constant phase difference over a wide frequency band when compared with a suitably chosen length $l_m$ of an ordinary microstripline. The differential phase shift can be controlled by the coupling factor between the top layer and the bottom layer for both sections [17]. This can be accomplished by adjusting the minor axis of the elliptical patches, $D_m$ and $D_c$ and slot, $D_s$.

The maximum return loss and minimum insertion loss over the operational frequency band can be achieved by tuning the length of the elliptical patches, $l_1$ and $l_2$ and slots, $l_3$, which have values close to quarter of the effective wavelength at the centre of the passband. After optimisation using the CST software, the dimensions of the phase shifter in mm are $D_m = 3.5$, $D_c = 1.5$, $D_s = 5.8$, $l_1 = 4.3$, $l_2 = 4.1$ and $l_3 = 4.9$, $l_m = 18.6$.

The manufactured quadrature Wilkinson divider is shown in Fig. 4b. For the substrate RT6010, the fabricated quadrature divider is of dimensions 28 × 24 mm excluding the connectors and thus represents a compact design in the intended frequency band. Fig. 5 shows the simulated and measured results of the designed device.

The return loss is better than 10 dB for both simulated and measured results across the band from 3 to 9.8 GHz. The simulated and measured isolation is greater than 15 dB across the whole band. As shown in Fig. 5a, a good agreement between the simulated and measured results can be noticed in the differential phase shift of the power divider from 3 to 10 GHz. In this frequency range the device achieves the differential phase shift of $90 \pm 7^\circ$. According to the simulated and experimental results presented in this section, each component exhibits the required characteristics to form a wideband six-port network.

**Fig. 5** Simulated and measured S-parameters of double-stage quadrature Wilkinson power divider

a. Return loss and isolation
b. Insertion loss
c. Differential phase shift

**Fig. 6** Fabricated six-port network

a. Top view
b. Bottom view
3 Results of the integrated six-port network

Fig. 6 shows the photo of the fabricated six-port network formed by in-phase and quadrature Wilkinson power dividers realising the structure of Fig. 1b. From Fig. 6, it can be seen that relatively long sections of transmission lines are used to connect the different devices forming the six-port network to accommodate, within reasonable distances, the six sub-miniature-A connectors, especially the two internal connectors shown in Fig. 6b, needed for the experimental testing.

The simulated and measured transmission coefficients from port 1 to output ports 3–6 of the designed six-port network are shown in Fig. 7.

Ideally, the values of these transmission coefficients should be −9 dB. As observed, the obtained simulated values are $\sim -10.5 \pm 1.5$ dB across 3–9.5 GHz. The measured transmission coefficient values from port 1 to ports 3–6 are $\sim -11 \pm 2$ dB across the frequency band from 3 to 9 GHz.

Similar results were obtained when port 2 is designated as input port, thus the transmission coefficient from port 2 to ports 3–6 are not shown in Fig. 7.

Fig. 8 illustrates the simulated and measured reflection coefficient at input port 1 and isolation between ports 1 and 2 of the proposed six-port network.

The simulated and measured reflection coefficient at port 1 is better than 10 dB over 3–10 GHz. Similar values for the reflection coefficient at port 2 are obtained. The simulated and measured isolation between ports 1 and 2 is better than 15 dB across the band from 3 to 11 GHz.

Fig. 9 shows the simulated and measured phase characteristics of the proposed six-port network. Theoretically the phase differences between ports 3 and 5 should be +90° and −90° between ports 6 and 4 when referring to input ports 1 and 2. The simulated results show almost a constant phase shift for the intended frequency band of 3–9 GHz where the phase imbalance of 90°/−90° is $\pm 15°$. The measured results show almost a constant phase shift for the intended frequency band of 3–8 GHz where the phase imbalance of 90°/−90° is $\pm 25°$.

The observed discrepancies between the simulated and measured results, especially in the phase performance that is very sensitive to any errors, can be explained by the use...
of coaxial to microstrip transitions in the measurement system that were not included in the simulations. The performance of the connectors used in this case is degraded at the upper end of the investigated frequency band. The other reasons for the discrepancies can be attributed to the manufacturing imperfections of the utilised chip resistors and the soldering imperfections of the chip resistors and port connectors.

From the presented full electromagnetic simulations and measured performance, it is apparent that the relative phase differences and transmission coefficients performances are degraded above 8 GHz. Therefore, the overall performance of the designed six-port network can be stated as more than one octave bandwidth from 3 to 8 GHz. These results are considered inspiring and better than the ones presented in [8]. The proposed six port formed by in-phase and quadrature Wilkinson dividers/combiners indeed delivers wideband performance, as initially planned in [8].

4 Conclusion

The design of a wideband six-port network formed by in-phase and quadrature Wilkinson power dividers has been presented. The design complements a previous work which shows that six-port networks aimed for reflectometers of measurements systems or quadrature phase shift keying (QPSK) modulators/demodulators in communication systems can be entirely formed by divider/combiners and 90° phase shifters. The accomplished work provides the missing solution for wideband 90° phase shifters, which are required to realise wideband quadrature dividers constituting the six-port networks. The wideband 90° phase shifters are realised using a two-section of broadside-coupled microstrip-CPW technology. Such phase shifters are compatible with microstrip Wilkinson dividers enabling the full integration of the six-port network in a single-layer substrate. This design has the advantage of eliminating one terminated load and it can be implemented on a single layer board. It has been proved via full-wave simulations as well as measurements that the designed device features a wide bandwidth in terms of return loss, transmission coefficient and phase shift characteristics. Its good performance indeed makes it suitable for the use in wideband measurement systems or transceivers employing QPSK modulation/demodulation.

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6 References

Planar Bandpass Filters for Ultra-Wideband Applications

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Abstract—The design of planar bandpass filters (BPFs) with ultra-wideband (UWB) behavior is presented. The proposed filters utilize broadband coupling between elliptical-shaped microstrip patches at the top and bottom layer of the filter’s structure via an elliptical slot located at the mid layer, which contains the ground plane. A theoretical model is presented to explain performance of the suggested filters. Results of calculation show that the elliptical microstrip-slot–microstrip broadband coupled structure can be used to build UWB BPFs with a flat group delay, which makes the presented configuration a good candidate for very narrow pulse transmission/reception. A design procedure for multisection broadband coupled filters is explained. BPFs, which use different sections of the proposed structure, were designed and manufactured. Results of simulation and measurement show that the developed devices have a 3-dB insertion loss bandwidth from 3.1 to 10.6 GHz. They have less than 1-dB insertion loss at the center of the passband, a sharp cutoff stopband, especially at the low-frequency stopband, and a flat group delay within the passband.

Index Terms—Bandpass filter (BPF), broadband coupling, ultra-wideband (UWB).

I. INTRODUCTION

THE RELEASE of the unlicensed use of the ultra-wideband (UWB) (3.1–10.6 GHz) by the Federal Communications Commission (FCC) has triggered significant research activities in both academy and industry toward exploring various UWB components and devices. Among them is the bandpass filter (BPF), which is a key device in almost any communication system. The BPF to be used in UWB systems is required to have low insertion loss over the band from 3.1 to 10.6 GHz, and a flat group delay performance within that band. Moreover, the UWB BPF should exhibit a very good selectivity in order to meet the FCC spectrum mask, especially at the cutoff region below 3.1 GHz. Developing a UWB-BPF is definitely a challenge. The first difficulty comes from the 110% fractional bandwidth requirement, which makes some widely used techniques for BPF design inapplicable. The second challenge comes from the requirement of a flat group delay over the UWB, and the third is the FCC limit at the low-frequency end.

It is well known that the microwave filter theory was established under the assumption of a narrow fractional bandwidth. As such, this theory is not applicable for the design of UWB filters, due to the unexpected spurious resonances, and uncontrollable nonlinear frequency dispersion over the 110% band.

The initial planar BPFs were designed using end-coupled coplanar waveguides (CPWs) [1], [2]. Due to the extremely weak coupling extent of these end-coupling elements, the resultant CPW filters suffer from a narrow passband and a high insertion loss. To circumvent this issue, various CPW edge- or parallel-coupled structures were suggested in [3] and applied in [4] to develop a wideband CPW filter with approximately 60% fractional bandwidth.

BPFs based on the combination of CPW low- and high-pass periodic structures were presented in [5] and [6]. These wideband filters have good suppression of out-of-band response. However, they have the drawback of a large circuit size, and an imperfect group delay over the passband. To meet the UWB filter specification, several studies to increase the number of sections have been reported [7], [8].

The parallel-coupled microstrip line with a slotted ground plane was employed to give the required tight coupling for a wideband BPF [9], [10]. However, increasing the fractional bandwidth requires the use of a very narrow, and sometimes impractical, gap size in spite of the modified three coupled-lines method presented in [11]. The BPF presented in [12] combined the effect of the parallel coupled lines with five short-circuited stubs. The use of many grounding vias complicated the manufacturing process, especially if we know that performance of the filter is very sensitive to size and position of the vias.

An initial UWB filter was constructed by mounting a microstrip line in a lossy composite substrate so as to attenuate the signals at high frequencies [13]. The reported performance of the filter shows more than 6-dB insertion loss in the passband. In another approach, a BPF with 49% fractional bandwidth was designed in [14] by using two stopbands of a filter block with two tuning stubs on a ring.

A compact UWB BPF on microstrip line was reported in [15] and [16] using a single multimode resonator, which is driven at two sides by two identical parallel-coupled lines. The idea of the multimode resonator UWB filter was proposed in [9] and applied in [17] to achieve a fractional bandwidth of 60%–80% with a stepped-impedance resonator [18]. The design presented in [16] has an insufficiently tight coupling between the parallel-coupled microstrip lines. To address the problems of the design in [16], a hybrid microstrip/CPW structure with a multimode resonator was presented in [19], whereas [20] suggested a BPF by forming a CPW-based multimode resonance in which two sides are linked with two single microstrip-to-CPW transition structures. In another approach, a UWB BPF with a short-circuited CPW multimode resonator was designed [21]. This type...
of filter can cover the UWB frequency band with low insertion loss. However, it has a narrow stopband at the high frequency.

A modified class of multimode resonator-based wideband filters was presented in [22]–[25] to improve the out-of-band rejection skirts. However, these filters can hardly achieve the UWB passband. An open-ended CPW multimode resonator with one low-impedance section in the middle and two high-impedance sections in two sides was used in [26] to construct a UWB BPF. Tight coupling and, hence, very low gaps were needed to achieve a UWB performance.

A UWB BPF was developed in [27] by adopting the high-pass filter prototype and transition stretch stubs to create the lower and upper stopbands. In [28], an electromagnetic bandgap periodic structure was utilized for the design of ultra-wide BPFs with harmonic passband suppression. The measured performance of the designed filter shows that it cannot cover the FCC-defined UWB passband.

The developed UWB BPFs mentioned above are mainly based on the traditional parallel- or edge-coupled line structure, in which a very strong coupling structure will be a must for a UWB performance. The tolerance of the microstrip and CPW fabrication process, however, imposes an upper limit upon coupling levels for parallel- and edge-coupled structures. This makes the production of such filters difficult using PCB technology as their performance is very sensitive to the production errors. This difficulty can be circumvented by implementing such high coupling using broadside coupling. In [29] and [30], the author used an elliptical-shaped broadside coupled structure via an elliptical slot to construct UWB couplers and phase shifters. In this paper, the broadside coupled multilayer structure is proposed as one of the possible solutions to build low-cost and efficient UWB BPFs.

This paper starts by presenting a theoretical model to analyze the proposed broadside-coupled microstrip-slot–microstrip structure. A design procedure for multisection BPFs is then explained. The measured and simulated results for five BPFs designed using the suggested method are presented. The designed BPFs show UWB behavior with flat group delay across the passband. They have also a wide rejected out-of-band with sharp cutoff, especially at the low-frequency band, where the FCC cutoff conditions are very strict.

II. ANALYSIS

The proposed BPF configuration is shown in Fig. 1. It utilizes broadside-coupled multilayer structure. Two similar elliptical-shaped microstrip patches at the top and bottom layers are coupled via an elliptical slot at the mid layer of the structure, which also contains the ground plane. The reason for the choice of this configuration is that the tapered shape of the elliptical broadside-coupled structure provides an almost constant tight coupling across the UWB, as proven in [29] and [30], where the same structure was used to build directional couplers and phase shifters.

Assume the device is designed to have a coupling equal to $\gamma$ between the top and bottom patches. Ports 1 and 2 are assumed to be the input and output ports respectively, whereas ports 3 and 4 are terminated in an open circuit. Depending on the odd- and even-mode analysis of the four-port devices, and assuming that the input and output ports are perfectly matched, the reflection coefficient at the input port ($S_{11}$), and the insertion loss from the input to the output port ($S_{21}$) can be calculated as follows [31]–[34]:

$$S_{11} = \frac{1 - \gamma^2 \left[1 + \sin^2(\theta f l)\right]}{\sqrt{1 - \gamma^2 \cos^2(\theta f l) + j \sin(\theta f l)}}$$

$$S_{21} = \frac{j \gamma \sqrt{1 - \gamma^2 \sin^2(\theta f l) + j \sin(\theta f l)}}{\sqrt{1 - \gamma^2 \cos^2(\theta f l) + j \sin(\theta f l)}}$$

where $l$ is the physical length of the coupled structure, and $\theta f$ is the effective phase constant in the medium of the coupled structure. In all the calculations that follows, the length $l$ is chosen such that $\theta f l = \pi/2$ at the center of the passband (6.85 GHz).

The group delay of the filter ($\tau$) is a function of the phase insertion loss. Applying the method presented in [35] on (2), $\tau$ can be found to be

$$\tau = \frac{2 \pi}{\Omega_0} \sqrt{\frac{\cos^2(\theta f l) \left[1 + \sin^2(\theta f l)\right]}{\cos^2(\theta f l) - \gamma^2}}$$

where $\Omega_0$ is the speed of light in free space and $\varepsilon_r$ is the dielectric constant of the substrate.

Assume that in order to increase the sharpness of the insertion loss decay at the stopbands below 3.1 and above 10.6 GHz, it is required to use $n$ sections of the broadside coupled structure shown in Fig. 1. In this case, the cascaded connection is performed as shown in Fig. 2.

It is interesting to note that, with an even number of sections, the input and output ports are located at the same layer, whereas with odd number of sections, the two ports are at different layers. In addition to the other factors related to the required performance, this could be an important parameter in deciding the number of sections to be used. For example, in the low-temperature co-fired ceramics applications, it is usually preferred to have the input/output ports of the filter at different layers. In this case, an odd number of sections is a must to avoid using an additional lossy transition.

In the cascaded connection shown in Fig. 2, an additional 50-$\Omega$ microstrip line with length $t_d$ is used to connect the two
consecutive sections in order to minimize the undesired mutual coupling between the top (or bottom) layers of those sections. The optimum value for $l_3$ is to be found after calculating the equivalent return loss and insertion loss of the entire structure. Performance of the cascaded structure can be estimated depending on the $S$-parameters of the different sections. The analysis can be started with a two-section filter, and the result can be generalized to any number of sections. In Fig. 3, a schematic diagram shows the incident ($b_1$) and reflected ($b_2$) signals at the $j$th port of the $j$th section. The midsection transmission line shown in Fig. 3 is assumed to be lossless and perfectly matched with the two sections of the BPF. Hence, it only introduces a phase shift equals to $e^{j2\pi/\lambda}$ on the signals $b_1$ and $b_2$. $\lambda$ is the phase constant of the microstrip line, which can be calculated from the well-known microstrip design equations [35].

Using the definition of the $S$-parameters on Fig. 3, it is possible to show that the return loss of the multisection filter ($S_{21} = b_2/b_1$) and its insertion loss ($S_{21} = |b_2/b_1|$) as a function of the $S$-parameters of its sections are equal to

$$S_{11} = S_{11}^{(n-1)} + \left(\frac{S_{21}^{(n-1)} e^{-j2\pi/\lambda}}{1 - S_{11}^{(n-1)} e^{-j2\pi/\lambda}}\right) e^{-j2\pi/\lambda} \frac{l_n}{1 - S_{11}^{(n-1)} e^{-j2\pi/\lambda}}. \quad (4)$$

In deriving (4), it was assumed that the filter is perfectly matched with the input and the output, and that the structure is reciprocal and symmetrical, i.e., $S_{11} = S_{22}$ and $S_{12} = S_{21}$.

The group delay ($\tau$) for the multisection BPF can be calculated by applying the method in [35] to (4).

Equation (4) can be generalized for $n$ cascaded sections as follows. The effective $S$-parameters for the $n$-section BPF can be calculated from the $S$-parameters of the first ($n-1$) sections ($S_{11}^{(n-1)}, S_{21}^{(n-1)}$) and the $S$-parameters of the last section ($S_{11}^{(n)}, S_{21}^{(n)}$). Using (4) and assuming that the $S$-parameters of the first stage in the equivalent two-section circuit is the effective $S$-parameters of the first ($n-1$) sections, it is possible to show that

$$S_{11} = S_{11}^{(n-1)} + \left(\frac{S_{21}^{(n-1)} e^{-j2\pi/\lambda}}{1 - S_{11}^{(n-1)} e^{-j2\pi/\lambda}}\right) e^{-j2\pi/\lambda} \frac{l_n}{1 - S_{11}^{(n-1)} e^{-j2\pi/\lambda}}. \quad (5)$$

$$S_{21} = \frac{S_{21}^{(n-1)} e^{-j2\pi/\lambda}}{1 - S_{11}^{(n-1)} e^{-j2\pi/\lambda}} + \frac{l_n}{1 - S_{11}^{(n-1)} e^{-j2\pi/\lambda}}. \quad (6)$$

This procedure can be repeated for the ($n-1$) sections, and so on, until the first two sections, where it is possible to use (4).

At this stage of the analysis, it is required to find the optimum length ($l_3$) for the additional transmission line connecting any two consecutive sections of the BPF. Using (4), the effect of $l_3$ on the return loss and insertion loss of a two-section filter is shown in Fig. 4, assuming different values for $l_3$. In these results, the coupling $\gamma$ is assumed to be equal to 0.7 for each of the two sections. It is clear from Fig. 4 that the increasing value of $l_3$ increases the 3-dB insertion-loss bandwidth of the filter at the high-frequency range. It has a negligible effect on the performance of the filter at the low-frequency range. However, increasing $l_3$ has a negative impact on the return loss at the passband. The analysis was repeated using a different number of sections and coupling factors. It was noticed that if $l_3$ is less than $\lambda_d/\sqrt{k}$, then it has a negligible effect on the performance of the multisection filter. $\lambda_d$ is the microstrip wavelength at the center frequency. If it is required to use $l_3$ to increase the fractional bandwidth without causing a significant negative impact on the passband performance, $l_3$ can be made as large as $\lambda_d/\sqrt{k}$.

Variation of the calculated $S_{11}$, $S_{21}$, and $\tau$ with frequency for one- to five-section filters are shown in Figs. 5 and 6 after using (1)–(6). In these calculations, it is assumed that all sections of the filter have the same coupling value. It is clear from Fig. 5 that it is possible to design a BPF with 110% fractional bandwidth by just using one section of the suggested structure. If it
is required to have a sharp cutoff and a higher insertion loss at the stopband regions with a wide high stopband, a multisection BPF is to be used. An increasing number of the filter’s sections, increases sharpness of the insertion loss decay at the cutoff regions, improves the performance at the high stopband, and decreases the fractional bandwidth with a little negative effect on the passband performance.

The proposed BPF has a flat group delay, as shown in Fig. 6. The lowest average group delay (≈ 0.05 ns) can be obtained using a one-section filter with a high coupling value (≈ 0.8). It is obvious from Fig. 6 that the group delay of the \(n\)-section filter (\(\tau_n\)) at the passband is approximately equal to \(n\) times the group delay of a one-section filter. This conclusion can be verified using (5) and (6). Within the passband, the reflection coefficient of each section (∼ 1) is very low. Hence, the denominator of (6) is approximately equal to 1. Assuming the same coupling for the different sections and using \(l_n = l\), (6) can be approximated to

\[
S_{21} \approx l \cdot S_{21}^1, n \cdot S_{21}^1, \ldots, \tau_n = n\tau_1.
\]

Therefore, the phase variation from the input to the output of the \(n\)-section filter at the passband is \(\psi_{21} = n\psi_{21}^1\), from which the group delay of the \(n\)-section filter is

\[
\tau_n = n\tau_1.
\]

The fractional bandwidth, rate of cutoff at the stopband, and the average group delay and its standard deviation were calculated for a wide range of coupling values and number of sections. Summary of the results is shown in Figs. 7 and 8.

The general behavior of the fractional bandwidth presented in Fig. 7 indicates that it decreases with increasing number of sections. It is also clear from Fig. 7 that the 110% fractional bandwidth required for UWB applications can be achieved by using a significant range of coupling values and/or number of sections. On the other hand, the designed BPF should have a sharp rate of cutoff at the stopband. From Fig. 7, it is obvious that in order to have a sharp cutoff at the stopband regions, a multisection filter should be used.

The other important parameter, which shows the quality of the BPF, is the group delay. The calculated values for the average group delay and its standard deviation for the proposed filters are shown in Fig. 8.

The system designer is usually concerned with the standard deviation of the group delay, whereas the average value has no significant effect on the overall performance of the system. The designed BPF should have a low value for the standard deviation of the group delay, or its peak-to-peak variation. The results shown in Fig. 8 reveal that the proposed BPFs have, in general, low values for the standard deviation of the group delay and its average value. There is a direct relation between the average value of the group delay and number of filter’s sections, as proven earlier, whereas the average value of \(\tau_1\) and its standard deviation are inversely related to the coupling value. The conclusion here is to use the highest possible coupling value.

According to the results presented in Figs. 6–8, it seems that increasing the value of the coupling \(\zeta\) improves performance of the filter. Actually, this conclusion should be restricted for a certain range of \(\zeta\), as there is an ugly side to the picture: increasing...
\( \zeta \) beyond a certain limit causes deterioration in the performance at the passband. In Fig. 9, variation of the insertion loss and return loss of a two-section BPF is shown for different values of \( \zeta \). It is obvious from the presented results that increasing the coupling value from 0.6 to 0.7 increases the bandwidth (as expected in Fig. 7) without a significant effect on the passband performance. If \( \zeta \) is increased to more than 0.8, the passband performance of the filter starts to deteriorate. Therefore, the optimum value of the coupling for the BPF designed using the proposed approach is roughly between 0.6–0.8. This conclusion was verified, and proven to be correct, for any number of the filter’s sections. Here, the importance of the suggested broadside-coupled elliptical structure becomes clear. A tight coupling \( \zeta \) is required to build a UWB BPF with good performance, something that is difficult to achieve using edge- or parallel-coupled lines.

The final step in the analysis, before building the filters, is to check whether it is better to design the multisection filter with the same value of the coupling for its different sections, or it is better to design each section having a different \( \zeta \). The return loss and insertion loss of a two-section filter was calculated assuming first the same coupling \( \zeta \) for the two sections, and then a different value for each section: \( \zeta_1 \) for the first section and \( \zeta_2 \) for the second section. The result of calculation, shown in Fig. 10, reveals that the best performance can be achieved when the two sections have the same value of coupling. If the two sections are designed using a different value of \( \zeta \), the resultant fractional bandwidth is equal to that of the two-section filter using the same coupling \( \zeta \) when \( \zeta_1 + \zeta_2 = 2 \zeta \), but with an inferior passband performance. The calculations were repeated for the three-section filter assuming all the possible cases including the case of symmetrically identical coupling factors. The conclusion was the same as for the two-section filter: there are no clear benefits from using different coupling factors for the different sections. Therefore, any multisection filter presented in this paper has the same coupling factor for all of its sections.

### III. Design

According to Figs. 7–9, the design of a UWB BPF requires a compromise between the required large number of sections with a high coupling value, to achieve a sharp cutoff at the stopband, and the required low number of sections to have a high fractional bandwidth and a low fluctuation in the group delay. After defining the design priorities, the required value of the coupling \( \zeta \) can be found from Figs. 7 and 8. This value can then be used to calculate dimensions of the coupling structure as follows.

Depending on value of the coupling, the even (\( \mathcal{Z}_{e} \)) and odd (\( \mathcal{Z}_{o} \)) mode characteristic impedances for the coupled patches are calculated using the following equations:

\[
\mathcal{Z}_{e} = \mathcal{Z}_{0} \sqrt{(1 + \zeta)/(1 - \zeta)} \quad \mathcal{Z}_{o} = \mathcal{Z}_{0} \sqrt{(1 - \zeta)/(1 + \zeta)}
\]

(9)

where \( \mathcal{Z}_{0} = \mathcal{Z}_{c} + j \mathcal{Z}_{l} \) is the characteristic impedance of the microstrip input/output ports of the device. Dimensions of the coupled region to give the impedance values calculated from (9) can be found by using the conformal-mapping technique [36]

\[
\mathcal{Z}_{e} = \frac{\sqrt{\mathcal{K}^{2}(\zeta_{1})}}{\sqrt{\mathcal{K}^{2}(\zeta_{1}) + \mathcal{K}^{2}(\zeta_{3})}}, \quad \mathcal{Z}_{o} = \frac{\sqrt{\mathcal{K}^{2}(\zeta_{1})}}{\sqrt{\mathcal{K}^{2}(\zeta_{1}) + \mathcal{K}^{2}(\zeta_{3})}}
\]

(10)

where \( \mathcal{K}(\zeta) \) is the first kind elliptical integral and \( \mathcal{K}^{2}(\zeta) = \mathcal{K}^{2}(\sqrt{1 - \zeta^{2}}) \). Major diameters of the elliptical coupled microstrip patches at the top and bottom layers (\( \Gamma_{1} \)) and slot (\( \Gamma_{2} \)) at the mid-layer can be found by using the conformal-mapping technique [36] and the method given in [29]

\[
l_{1} = \frac{h}{\sqrt{\sin \lambda_{c} (\pi^{2} \Gamma_{1} J/(4 h \lambda_{c}))}}
\]

(11)

\[
l_{2} = \frac{h}{\sqrt{\sin \lambda_{c} (\pi^{2} \Gamma_{2} J/(4 h \lambda_{c}))}}
\]

(12)

\[
l_{3} = \frac{h}{\sqrt{\sin \lambda_{c} (\pi^{2} \Gamma_{3} J/(2 h \lambda_{c}))}}
\]

(13)

where \( \lambda_{c} \) is the wavelength inside the substrate at the center frequency (6.85 GHz), \( h \) is the thickness of the substrate, \( h_{0} \) is the
TABLE I
VALUES OF THE DESIGN PARAMETERS

<table>
<thead>
<tr>
<th>Number of sections</th>
<th>( C )</th>
<th>( \epsilon_r )</th>
<th>( h )</th>
<th>( f_0 )</th>
<th>( A )</th>
<th>( B )</th>
<th>( D )</th>
<th>( l_0 )</th>
<th>( l_1 )</th>
<th>( l_2 )</th>
<th>( l_3 )</th>
<th>( A )</th>
</tr>
</thead>
</table>

\[ l = \left( \sqrt{\left(\frac{\lambda_c}{4}\right)^2 + \left[\pi \Gamma_{\text{opt}} l / \lambda_c \right]^2 + \lambda_c / 4 \right)^2 / 2 \]  

It is to be noted that if the device is not put in an enclosure, then \( h_0 \) in (14) can be assumed to be infinitely high. Hence, from (13), \( l_3 = \frac{1}{2} \text{ and } (10) \text{ reduces to the equation given in [29].} \)

The final step in the design procedure is to find width \( b \) of the input/output microstrip lines, which connects the filter to the 50-\( \Omega \) input/output ports. This can be achieved using standard microstrip design equations [35].

Dimensions of the BPFs calculated using the proposed method and optimized using Ansoft HFSSv10 software are shown in Table I. The optimization was aimed to improve performance of the filters at the high stopband (above 10.6 GHz). The devices were assumed to use Rogers RT6010LM as a substrate. No shielding box was considered and, therefore, value of \( l_3 \) in (10) was assumed to be 0. For the multisection filters, the same coupling value was considered for the different sections. It is worth mentioning that there is approximately a 5% difference between the calculated and optimized values of \( \Gamma_{\text{opt}}, l \), and \( \Gamma_{\text{opt}} \), whereas the optimized value for \( \Gamma_{\text{opt}} \) is less than the calculated value by around 15% for all the designed filters.

It is to be noted from Table I that the coupling value for the one-section filter was taken to be 0.68, although it is clear from Fig. 7 that 0.6 coupling can achieve over 110% fractional bandwidth. This is because the calculated return loss, when \( \Gamma = 0 \), fluctuates at the passband between 12–8 dB, whereas it is better than 20 dB across the passband when \( \Gamma = \text{optimized} \).

IV. RESULTS

The five BPFs, designed as in Section III, were manufactured using Rogers RT6010LM (with \( \epsilon_r = 10.2 \), thickness = 0.0084 mm, and tangent loss = 0.0012) as the substrate. The overall dimensions of these devices are 15 mm × 20 mm, 15 mm × 25 mm, 15 mm × 30 mm, 25 mm × 30 mm, and 25 mm × 30 mm for the one-, two-, three-, four-, and five-section filters, respectively. These dimensions confirm the compact size of the developed devices. A photograph for the developed five-section filter is shown in Fig. 11. This photograph reveals how the space was efficiently utilized by distributing the filter’s sections around an arc. The same design strategy was also used with the four-section filter.

The developed devices were tested via simulations and measurements. The simulation was done using the full electromagnetic analysis software Ansoft HFSSv10, whereas the measurements were done using a vector network analyzer. Fig. 12 shows the simulated and measured results for the three- and five-section BPFs. Due to the space limitation, the results for the one-,
two-, and four-section filters are not shown, but their performances are explained here. The measured and simulated results indicate that the manufactured one-, two, and three-section filters have a passband that covers the FCC-defined UWB, whereas the four- and five-section filters cover the band from 3.2 to 10 GHz. The insertion loss of the designed filters at the center of the passband is less than 1 dB, whereas the return loss is higher than 25, 17, 12, and 11 dB for the one-, two-, three-, four-, and five-section filters, respectively. Sharpness of the cutoff stopband region increases with the increasing number of the filter’s sections. The overall performance of the filters agrees well with the predicted ones using the theoretical model. There is also a good agreement between the measured and simulated results.

The simulated and measured results reveal that part of the high stopband suffers from the effect of a spurious response. With the help of the theoretical model presented in Section II, it is possible to show that the center of the spurious response appears at a frequency that is an odd multiple of the center frequency of operation. Hence, the first peak of the spurious response appears at 20.5 GHz, which is far away from the UWB, permitting a relatively wide stopband after 10.6 GHz.

The measured insertion loss of the four- and five-section filters across the high-frequency portion of the passband is higher than the calculated and simulated value. The measured insertion loss was found to be over 2 dB at the frequency range from 8 to 10 GHz. The calculated value is less than 1 dB for that range. This increase in the passband insertion loss is mainly due to the radiation losses from the microstrip patches at the top and bottom layers in addition to the insertion loss of the subminiature A (SMA) connectors used in the manufactured filters.

To eliminate effect of the radiation losses, the four- and five-section filters were redesigned so that they can be enclosed within a shielding box. The dimension of the box is 25 mm × 30 mm × 20 mm, which means that \( b_{o} = \frac{2a}{b} \text{mm} \) in (13). It was noticed that with the shielding box, \( r_{e} \), for the new filters is 5% higher, whereas \( r_{g} \) is 5% lower than the values of the unshielded filters. The redesigned filters were tested. Results of the measurement, which are shown in Fig. 12, indicate that the insertion loss at the high-frequency portion of the passband was improved significantly. The redesigned filters have an insertion loss that is less than 1.2 and 1.7 dB for the four- and five-section filter, respectively, across their passband.

The group delay of the five manufactured filters were simulated and measured. The results for two of the developed filters are shown in Fig. 12. The filters have a flat group delay across the UWB. The average measured group delay for the filters is 0.2, 0.35, 0.42, 0.57, and 0.72 ns, whereas the peak-to-peak variation in the group delay within the passband of the filters is 0.04, 0.06, 0.1, 0.2, and 0.25 ns for the one-, two-, three-, four-, and five-section filter, respectively. There is a good agreement between the measured and simulated results for the group delay of the designed filters.

There is a significant difference between the average measured and simulated values of the group delay presented in Fig. 12 and the calculated values shown in Fig. 6. This is due to the effect of the additional transmission lines used to connect the filter to the input/output SMA ports and the additional line between different sections of the devices. The results introduced in Fig. 6 are only for the effect of the coupled structure. The effect of the additional transmission lines on the average group delay can be calculated as follows. The value of \( S_{21} \) for the transmission line of length \( l_{a} \), is \( e^{-j \frac{2\pi l_{a}}{\lambda}} \). Using the definition of the group delay \( \tau \), the additional \( \tau \) due to the transmission line is \( \frac{\lambda}{c} \frac{\partial S_{21}}{\partial \tau} \), where \( \varepsilon_{r} \) is the effective dielectric constant and can be calculated using the microstrip design equations [35]. Substituting for the length \( l_{a} \), used in the manufactured filters gives the following additional group delay: 0.12, 0.15, 0.12, 0.22, and 0.2 ns for the one-, two-, three-, four-, and five-section filters, respectively. Subtracting these numbers from the measured magnitudes in Fig. 12 gives average group delay values, which are close to those calculated and shown in Fig. 6.

V. CONCLUSION

The design of planar BPFs with UWB behavior has been presented. The proposed filters utilize broadside coupling between elliptical-shaped microstrip patches at the top and bottom layer of the filter’s structure via an elliptical slot located at the mid layer, which contains the ground plane. A theoretical model has been presented to explain performance of the presented configuration. Results of the calculation have shown that the elliptical microstrip-slot–microstrip broadside-coupled structure can be used to build UWB BPFs with a flat group delay, which makes the presented design a good candidate for UWB pulse transmission/reception. A design procedure for multisection broadside-coupled BPFs has been explained. BPFs of different sections of the proposed structure were designed and manufactured. Results of simulation and measurement have shown that the developed devices have a 3-dB insertion loss bandwidth from 3.1 to 10.6 GHz. They have less than 1-dB insertion loss at center of the passband, a sharp cutoff stopband, especially at the low-frequency stopband, and a flat group delay within the passband.

The suggested configuration offers another advantage, which is the possibility of the direct integration in the multimode microstrip/slot technology without using additional transitions or external connectors. The presented design should be of special interest to the designers of modern multilayer microwave circuits, such as the low-temperature co-fired ceramics and the laminated multichip modules. This is because they efficiently tackle the problem of UWB signal distribution across different layers.

REFERENCES


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Multilayer bandstop filter for ultra wideband systems

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Abstract: A multilayer broadside-coupled microstrip-slot-microstrip structure is used to design a bandstop filter with a wide passband for ultra wideband (UWB) applications. The design procedure for the proposed filter is based on the conformal mapping technique and the even- and odd-mode analysis. The theoretical analysis indicates that value of the coupling factor between the top and bottom layers of the structure can be used to control the width of the stopband, whereas centre of that band can be controlled by the length of the coupled structure. To limit the passband of the proposed bandstop filter to 3.1–10.6 GHz, which is the specified bandwidth for UWB systems, a broadside-coupled bandpass filter is integrated with the proposed device. The simulated and measured results show that the proposed device achieve <0.5 dB insertion loss across most of the passband and >20 dB insertion loss at the stopband. The device also shows a flat group delay across the passband with <0.15 ns peak-to-peak variation. Hence, it is a suitable choice for the UWB systems that require a distortionless operation.

1 Introduction

Bandstop filters (BSFs) play an important role in the modern communication systems because they are required to remove the undesired spurious components, such as harmonics, parasitic effects and intermodulation distortion through a system. BSFs should offer high levels of rejection over the undesired band, while creating minimal disturbance in the remainder of the spectrum.

Recent years have witnessed an increased interest in the ultra wideband (UWB) systems which cover the band 3.1–10.6 GHz. The UWB systems need to meet the condition of a harmonious operation with the existing standards which cover part of the assigned UWB frequencies, such as the IEEE802.11 and HIPERLAN systems which operate within the band 5–6 GHz. Therefore a UWB system which is expected to coexist with those systems should be equipped with a filter having a 3.1–10.6 GHz passband and a stopband which covers the frequencies used by the other coexisting systems.

The conventional method to design a BSF is to use open-circuited or short-circuited stub resonators which are electrically, or magnetically, coupled to the main transmission line. This method has been extensively used by the researchers and it generally results in a narrow stopband [1–9].

The spurline BSF, which belongs to the resonant-type filters with the resonator embedded within the main transmission line, was investigated by several authors [10–13]. The spurline filters have compact structures with a significantly low radiation loss compared with the conventional shunt stub filters. In [14, 15], microstrip BSFs using shunt open stubs and spurlines were presented. These kinds of BSFs have relatively narrow bandwidths because of the reliance on a resonance structure. Moreover, the resonant-type BSFs have the first spurious stopband at a frequency which is three times that of the fundamental stopband centre frequency. A solution to this problem was proposed in [16] using circuit optimisation. The filter structure presented in [16] was redesigned in [17] using a systematic circuit-oriented approach which provides an insight into the physical operation of these filters and requires minimal circuit optimisation.
The other trend in the design of BSFs is to utilise the non-resonant types of BSF such as the parallel-coupled transmission lines, which were theoretically proven to be able to achieve a broadband BSF [18]. However, it was shown that fabricating a wideband BSF by using the conventional-coupled resonators is difficult because of the need for very close spacings. In an alternative approach, cross-over and broadside-coupled lines were adopted to design BSFs [19]. However, the results obtained in [19] indicate a serious problem in the performance; the insertion and return losses have a very slow variation from their values at the stopband to the required values at the passband. In [20], a strong coupling was achieved by utilising a double negative-coupled transmission line, whereas the design presented in [21] relaxed the requirement of a narrow spacing by utilising a combination of loosely parallel-coupled input/output ports and a standard multiple-stub BSF.

For wide stopband BSFs, electromagnetic bandgap structures as well as slotted ground structures have been widely applied to microwave systems [22–30]. In these structures, the stopband is dependent on the shape, position and size of the etching in ground plane, which results in a large dimension and alignment problem between the signal strip-line and ground plane. In order to avoid the etching in the ground plane, the structures with T-shaped stubs at two sides of the main line were proposed in [31].

The metamaterial technology has been utilised in the design of BSFs as another form of resonant type of filters [32, 33]. The split-ring resonator was used in [32] to construct filters that need to notch certain frequencies sharply. The split-ring resonator was originally introduced in [34] to construct the left-hand materials. It can be considered as an electronically small resonant structure with a high quality factor. In [33], a notch BSF, which was previously proposed [35], with a left-handed transmission line was presented. The effectiveness of the left-handed metamaterials in size reduction and performance enhancement was demonstrated.

The multilayer broadside-coupled structure has been recently used by the author to design UWB directional couplers [36], phase shifters [37] and bandpass filters [38]. It has been shown that a wide range of coupling values can be achieved using the multilayer broadside-coupled structure with a UWB performance. In this paper, a multilayer broadside-coupled microstrip-slot-microstrip structure is utilised to design a BSF which has a UWB passband and a certain stopband that can be controlled using the value of the coupling factor. The proposed design avoids the limitation in width of the stopband associated with the resonated filter structures and the fabrication difficulties in the case of the parallel-coupled filters. Moreover, the use of the broadside-coupled structure enables to design BSFs with wide stopband as it is easy to achieve a wide range of coupling values in contrast to the limited range of coupling values in the previous designs which uses the parallel-coupled structures.

2 Design

The proposed BSF consists of two microstrip patches which are located at the top and bottom layers of the structure. The two patches are broadside-coupled via a slot in the ground plane, which is located at the mid layer. The three layers of the device are shown in Fig. 1, whereas Fig. 2 shows the whole structure. To reduce the radiation losses, the proposed device is assumed to be enclosed by a shielding box. It is worthwhile to mention that the structure shown in Figs. 1 and 2 contains also a two-stage bandpass filter, which is needed to limit the lower and upper ends of the passband as explained later in the paper.

The design of the proposed BSF is accomplished using the odd–even analysis of the coupled microstrip lines. Following the design steps in [37], the structure of the BSF shown in Fig. 2 can be considered as a four port backward coupler with one of the output ports terminated in an open circuit, whereas another port is terminated in a short circuit, (Fig. 3). Assume that the device is designed to have a coupling factor equal to $C$ between the top and bottom patches and that the input and output signals to/from the $i$th port are $a_i$ and $b_i$, respectively. As port 3 is terminated in an open circuit, whereas port 4 is terminated in a short circuit, then the reflection coefficient at these ports is equal to 1 and $-1$, respectively. If the input port 1 and output port 2 are assumed to be perfectly matched then using the principles of the four ports backward coupler [39], the output signals at the four ports of the device can be

Figure 1 Layers of the proposed BSF integrated with a two-stage bandpass filter
a Top layer
b Mid layer
c Bottom layer

---

*Figure 1: Layers of the proposed BSF integrated with a two-stage bandpass filter.*

- **a** Top layer
- **b** Mid layer
- **c** Bottom layer
calculated as follows

\[ b_1 = Aa_3 \]  
\[ b_2 = Ba_1 + Aa_4 \]  
\[ b_3 = a_3 = Aa_1 + Ba_4 \]  
\[ b_4 = -a_4 = Ba_3 \]  

\[ A = \frac{jC \sin (\beta l_s)}{\sqrt{1 - C^2 \cos (\beta l_s) + j \sin (\beta l_s)}} \]  
\[ B = \frac{\sqrt{1 - C^2 \cos (\beta l_s) + j \sin (\beta l_s)}}{\sqrt{1 - C^2 \cos (\beta l_s) + j \sin (\beta l_s)}} \]

where \( l_s \) is the length of the coupled structure and \( \beta \) the effective phase constant in the medium of the coupled structure. For the structure under investigation, it is possible to show that

\[ \beta = \frac{\beta_e + \beta_o}{2} = \frac{2\pi\sqrt{\varepsilon}}{\lambda} \]

where \( \beta_e \) and \( \beta_o \) are the phase constant for the even and odd modes, respectively, \( \lambda \) the free space wavelength and \( \varepsilon \) the dielectric constant of the substrate.

From (1)–(4) and knowing that \( S_{11} = b_1/a_1 \) and \( S_{21} = b_2/a_4 \) are the return loss of the input port and the insertion loss from the input to the output port, respectively, then

\[ S_{11} = \frac{A^2}{1 + B^2} \]  
\[ S_{21} = B \left( 1 - \frac{A^2}{1 + B^2} \right) \]

To design a high-performance BSF, it is required to make the return loss \( (S_{11}) \) equal to 1 and the insertion loss \( (S_{21}) \) equal to zero at the centre of the stopband \( (f_r) \). It is possible to solve (8) and (9) under such a condition and to show that it can be achieved when the length of the coupled structure \( (l_s) \) is equal to the quarter of the effective wavelength calculated at \( f_r \) for any value of the coupling factor.

The design equations (8) and (9) can be used to show the effect of the coupling factor \( C \) on the performance of the proposed filter. The result is shown in Fig. 4, which reveals variation of the return and insertion losses of the device across the UWB. It is clear from Fig. 4 that width of the stopband depends on the value of the coupling factor. Increasing value of the coupling increases the width of the stopband and vice versa. This proves the need for a tight-coupled structure, such as the multilayer broadside-coupled configuration of this paper, to build a BSF with wide stopband.

It is to be noted from Fig. 4 that the passband of the filter extends well beyond the band assigned for UWB, which is 3.1–10.6 GHz. Therefore a bandpass filter is needed to limit the operation within the required band. The multilayer broadside-coupled structure presented in [38] can be used as it is compatible with the multilayer structure of the device presented in this paper (Fig. 2). It is valuable to mention that the coupled structure used for the BSF is of a rectangular shape, whereas it is of an elliptical shape for the bandpass filter (BPF). The reason behind that is the BSF is required to have a constant coupling factor only across the stopband, which is 5–6 GHz for the design presented in this paper. A simple rectangular shaped structure can achieve that goal. On the other hand, the BPF
is required to have a constant coupling factor across the whole UWB (3.1–10.6 GHz). Therefore there is a need to use some sort of a tapered structure, such as the elliptical-shaped structure, which has a constant coupling factor across the UWB [37]. If it is required to design a BSF with a very wide stopband then the elliptical shape is to be used for the coupled structure.

Performance of the BSF after including a two-section BPF designed according to the procedure given in [38] was calculated and it is shown in Fig. 5. The lower and upper ends of the passband of the integrated device have been limited by the added BPF so that they are 3.1 and 10.6 GHz, respectively.

To find the required dimension for this coupled structure, the odd- and even-mode analysis can be used. Depending on the required width of the stopband, the coupling factor $C$ can be estimated from Fig. 5. Using the estimated value of the coupling factor, the even ($Z_{oo}$) and odd ($Z_{oe}$) mode characteristic impedances for the coupled patches are calculated using the following equations

$$Z_{oo} = Z_0 \sqrt{\frac{1 + C}{1 - C}}$$
$$Z_{oe} = Z_0 \sqrt{\frac{1 - C}{1 + C}}$$

(10)

where $Z_0$ is the characteristic impedance of the microstrip ports of the coupler.

Assuming a quasi-transverse electromagnetic propagation, the electrical characteristics of the coupled lines can be completely determined from the effective per unit length capacitances and the phase velocity of the lines [40]. The characteristic impedance of each of the coupled lines at any of the two modes can be found using the relation

$$Z_{cm} = \frac{1}{v \sqrt{C_m C_{cm}}}$$

(11)

where $v$ is the velocity of light in free space, the subscript $m$ refers to the mode, $C_m$ is the per unit length capacitance of the coupled line at the $m$-mode and $C_{cm}$ is the capacitance when the dielectric is replaced by free space.

Using the conformal mapping techniques [41], the per unit length capacitances can be found as a function of the structure dimension and the substrate’s characteristics. Substituting the results in (11), the final results for the characteristic impedances of the coupled lines at the two modes of operation are

$$Z_{oo} = \frac{60 \pi}{\sqrt{E_i}} \left[ \frac{K'(k_s) + K(k_s)}{K(k_s)} \right]^{-1}$$
$$Z_{oe} = \frac{60 \pi}{\sqrt{E_i}} \left[ \frac{K(k_e) + K'(k_e)}{K(k_e)} \right]^{-1}$$

(12)

$$k_1 = \frac{\sinh^2 (\pi \omega_s / 4h)}{\sinh^2 (\pi \omega_s / 4h) + \cosh^2 (\pi \omega_s / 4h)}$$
$$k_2 = \tanh (\pi \omega_s / 2h_o)$$
$$k_3 = \tanh (\pi \omega_s / 4h)$$

(13)

where $\omega_s$ is the width of the slot in the ground plane, $w_c$ the width of the coupled lines at the top and bottom layers, $h$ the thickness of the substrate, respectively, $h_o$ the height of the enclosure and $K(k)$ the first kind elliptical integral and $K'(k) = K(\sqrt{1 - k^2})$.

It is to be noted that the direct relation between the coupling factor $C$ and the required dimensions of the coupled structure can be obtained by finding $C$ as a function of $Z_{oo}$ and $Z_{oe}$ from (10) and then substituting for these impedances from (12).

3 Results and discussions

The above method was used to design and build a BSF with a 5–6 GHz stopband and a passband which extends over the rest of the UWB frequency range. Rogers RO4003C with thickness 0.508 mm, dielectric constant 3.38 and tangent loss 0.0023 was used as a substrate. The design process was aided with a full electromagnetic simulation package (Ansoft HFSSv10), whereas the measurements were accomplished using a vector network analyzer.

According to Fig. 5, the coupling factor for the BSF should be around 0.4 to achieve a 5–6 GHz stopband, assuming the 3 dB insertion loss as a reference. Using $C = 0.4$, the even ($Z_{oe}$) and odd ($Z_{oo}$) mode characteristic impedances for the coupling patches can be found from (10) to be 76.4 and 32.7 Ω, respectively. Using the
A multilayer broadside-coupled structure has been used to design a BSF with a wide passband for UWB applications. The design method for the proposed filter is based on the conformal mapping technique and the even- and odd-mode analysis. The theoretical analysis has shown that width of the stopband has a direct proportional relationship with the coupling factor between the top and bottom layers of the structure, whereas centre of that band depends on the length of the coupled structure. To limit the passband of the proposed BSF to 3.1–10.6 GHz, a broadside-coupled bandpass filter has been integrated with the proposed device. The simulated and measured results have revealed that the proposed device has <0.5 dB insertion loss across most of the passband and >20 dB insertion loss at the stopband. The device has also shown a flat group delay across the passband with less than 0.15 ns peak-to-peak variation.

5 References


The designed divider is fabricated using microstrip lines on a Teflon substrate with the dielectric constant 2.1 and the thickness 0.5 mm. Figure 6 is the photograph of the fabricated divider.

The frequency response of the designed divider is shown in Figure 7. The measured results are compared with those from the full-wave electromagnetic simulation [15]. The measured data show reasonable agreement with the simulated results. The dual-band operation characteristic is clearly observed at the desired band frequencies 1 GHz and 2 GHz with 3 dB difference between the two outputs. The measured performances of the designed divider at the two band frequencies 1 GHz/2 GHz are: the input matching $-29.3\, \text{dB}/-20.3\, \text{dB}$ ($S_{11}$), the output matching $19.7\, \text{dB}/-18.9\, \text{dB}$ ($S_{21}$) and $-21.9\, \text{dB}/-26.5\, \text{dB}$ ($S_{32}$), the output isolation $-24.3\, \text{dB}/-32.2\, \text{dB}$ ($S_{23}$) and the output transmission $-4.93\, \text{dB}/-4.85\, \text{dB}$ ($S_{12}$) and $-1.92\, \text{dB}/-2.23\, \text{dB}$ ($S_{13}$).

4. CONCLUSIONS

A new Wilkinson power divider structure is proposed for the dual-band, unequal power division with matched output ports. The compact dual-band stubbed II sections are employed to reduce the overall divider length in half as compared with similar dividers based on the two-section cascaded sections. Compact open stub based design method also facilitates the low cost fabrication of the divider without using any lumped reactive elements. Exact design formulas are provided along with the detailed design procedure of the proposed divider, which are validated through the design and experiment of a prototype divider.

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passband suppression by employing electromagnetic bandgap periodic structures [9].

The concept of designing BPFs by connecting two identical tapered slot antennas (TSA) in the face-to-face configuration has recently been proposed by the authors [10]. In this article, that approach is extended to build BPFs with narrow and wide fractional bandwidths by utilizing the tapering profile of the employed TSA to control bandwidth of the filters. To improve its high frequency stopband performance of the BPF, the CPW-feeder of the filter is modified to include open-ended series stubs. The success of the proposed concept is demonstrated via full-wave electromagnetic simulations and measurements.

2. DESIGN

The antenna, which is utilized as a building block for the design of a wideband filter, is shown in Figure 1a. It belongs to a class of TSA which have attracted a significant attention due to their wideband performance, high directivity, low cost, and easy integration with the radio frequency circuitry [11]. An inherent feature of TSA is that it has a natural low cut-off frequency. This cut-off frequency is determined by the width \( w \) and depth \( d \) of the opening of the radiating aperture. It has been proven that the tapering profile of the tapered slot antenna has a direct effect on its directivity and bandwidth. The linearly tapered TSA has a wideband, whereas the concavely tapered antenna [see Fig. 1(c)] has a narrowband [12]. Therefore, two types of tapering were used to design both wideband and narrowband filters. The linearly tapered TSA used to design a wideband filter is shown in Figure 1(a), whereas the elliptically concaved TSA for a narrowband filter is shown in Figure 1(c). The antennas used in the proposed method are uniplanar with no ground plane at the bottom layer of the structure.

The feed structure of the two antennas includes a coplanar waveguide with the inner conductor tapered in the manner shown in Figure 1(a) to achieve a perfect matching across the whole band of interest.

Width of the opening of the antenna \( w \) was chosen to be equal to the effective wavelength at the lowest frequency of the passband, and the depth \( d \) was chosen to be a quarter of the effective wavelength at that frequency. The low cut-off frequency for the device was assumed to be 5 GHz to have a passband, which is located within the ultra wideband frequency range of 3.1–10.6 GHz. Assuming the use of Rogers RT6010 (\( \varepsilon_r = 10.2, \) thickness = 0.64 mm) as a substrate, the initial dimensions of the antenna were calculated to be: \( w = 30 \) mm and \( d = 7.5 \) mm. For the 50 \( \Omega \) CPW feeder, the dimensions were \( w_f = 2 \) mm and \( s = 0.5 \) mm.

To form a BPF, two identical TSA antennas are connected in the face-to-face orientation as depicted in Figures 1(b) and 1(d) to form a tapered slot resonator. The overall dimension of both filters is 25 mm \( \times \) 30 mm. To minimize the insertion loss of the designed filters at the passband, the structures were assumed to be enclosed in a metallic box. This configuration is used to prevent any radiation from the two antennas that can contribute to a higher insertion loss at the passband and to electromagnetic interference when these filters operate in presence of other RF circuits. Dimensions of the box were: width = 25 mm, length = 30 mm, and height = 20 mm. The filter board was located at center of the box.

The performance of the filter was optimized using the full-wave electromagnetic simulator CST Microwave Studio. The optimization was performed to achieve two main targets: Firstly to have a passband located within the desired frequency range with a minimum insertion loss, and secondly to have the widest possible higher stopband. The design parameters \( w \) and \( d \) for the filters after optimization were found to be 24 mm, and 7.7 mm, respectively, for the linearly tapered structure [Fig. 1(b)] and 21 mm and 10 mm for the elliptically concaved filter [Fig. 1(d)]. It is to be noted that when the shielding boxes were removed; the result was a small increase of about 1 dB in the insertion loss across the passband of the designed filters.

Figure 1 Linearly (a) and elliptically concaved (c) tapered slot antennas and the filters (b & d) formed from two of those antennas connected face-to-face to form a tapered resonator. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

Figure 2 Variation of the S-parameters with frequency for the designed filters. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
3. RESULTS AND DISCUSSIONS

The simulated performance of the two designed filters is shown in Figure 2. The linearly tapered filter has a wide passband that extends from 4.9 GHz to 8.2 GHz. The insertion loss at the center of the passband is less than 0.5 dB. The concavely tapered filter has a passband that extends from 4.3 GHz to 6.2 GHz with an insertion loss of less than 0.3 dB at center of the passband. Concerning their performance at the lower and upper stopbands, both filters show sharp low stopbands with an insertion loss of 33 dB at 1 GHz.

Performance of the two devices as BPFs can be explained by the simulated surface current distribution shown in Figure 3. At 1 GHz, which is well below the lower cut-off frequency of the filter, the input signal is completely reflected back at the input port and no current flows in the output port. At 5.2 GHz, which is within the passband of the filter, the input current flows via the input port and couples almost completely to the output port. This operation can also be explained by the passband characteristics of the TSA when operating in isolation. In the filter configuration, they are near field coupled. In the presented design, their residual radiation in the passband is minimized using a conducting enclosure.

The two filters exhibit a sharp cut-off at the beginning of the higher stopbands. The insertion loss is more than 40 dB at 10.2 GHz for the linearly tapered filter, while it is more than 20 dB at 8.3 GHz for the concaved filter. However, the two filters suffer from a spurious response at integer multiples of the center of their passbands. The linearly tapered filter demonstrates some resilience to that effect, while the concaved filter shows a deteriorated performance at its higher stopband due to those spurious responses.

To overcome the spurious response problem, the feeding structure of the two filters at the input and output ports was modified to accommodate open-ended series stubs. The modified structure behaves as a BPF with sharp high frequency stopband [13]. Length of each series stub was chosen initially to be half of the effective wavelength calculated at center of the spurious response. Because across the higher stopband, there are at least two significant spurious responses as revealed in the results of Figure 2, the modified feeder at the input port was designed to remove one of those spurious responses, whereas the feeder at the output port was designed to overcome the other. Those values were optimized to achieve the largest possible insertion loss across the stopband that extends up to 15 GHz.

The optimized structures of the two filters were manufactured. A photo of the devices is shown in Figure 4. In the inset of that photo, the modified feeder is shown. The optimized lengths of the series stubs \((l)\) are 4.6 mm and 4.1 mm for the linearly tapered filter, and 5.6 mm and 5.3 mm for the concaved filter. The width of the slots in the modified feeder was fixed at 0.1 mm, which is the lowest possible value that can be achieved using the milling machine available to the authors.

The two manufactured filters were tested using a vector network analyzer, and the results were compared with simulations. Figure 5 shows the simulated and measured results for the return
and insertion losses of the linearly tapered BPF. In the legend of the results, the subscript (m) refers to the measured values. The results shown in Figure 5 indicate that the device has a wide passband, which extends between 4.8 and 7.8 GHz according to the simulations and 4.8–8.2 GHz according to the measurements, assuming the 3 dB insertion loss as a reference. This result represents more than 50% fractional passband. The insertion loss at the center of the passband is less than 1 dB according to the measured results, whereas it is approximately 0.5 dB in the simulations. The filter has a sharp lower stopband with insertion loss of around 40 dB at 1 GHz. The effect of the modified feeder is revealed in the results of Figure 5. The spurious responses at the higher stopbands have been highly attenuated. The insertion loss is larger than 15 dB up to 14 GHz, where the insertion loss then starts to decrease till it becomes between 9 and 10 dB at around 15 GHz.

In the concaved filter, the simulated and measured results of Figure 6 reveal around 30% fractional bandwidth with the passband extending from 4.3 to 6.2 GHz in the simulations and 4.3 to 6.6 GHz in the measured results. The reduction in width of the passband is thought to result from the concaved tapering profile utilized for this filter. The insertion loss at the center of the passband is less than 0.2 dB in the simulations and less than 1 dB in the measurements. This discrepancy is expected, as the measured results include extra losses introduced by the CPW to coaxial transitions used in measurements. The low stopband (below 4 GHz) is sharp, providing an insertion loss of around 35–40 dB at 1 GHz. Performance of the filter at the higher stopband has been improved with the use of the modified feeder as seen from a comparison of the results in Figure 6 with those of Figure 2.

One other important characteristic of a BPF is a flat group delay across the passband. Concerning the manufactured filters, the group delay for the designed filter was measured and the results are depicted in Figure 7. It is clear from this result that the two developed filters have a group delay, which has a peak-to-peak variation of less than 0.5 ns across their passband. This result is an attractive feature for many applications.

4. CONCLUSION

The design of BPFs from two face-to-face connected planar TSA that form a tapered resonator has been presented. The tapering profile of the utilized antennas is used to control the fractional bandwidth. To improve the performance at the higher stopbands, the transmission feedline of the device was modified to include an open-ended series stub. The measured and simulated results of the filters have shown more than 50% fractional passband for the linearly tapered structure and around 30% fractional passband for the concaved structure, with less than 1 dB insertion loss at the center of the passband. The two filters have a sharp lower stopband performance, whereas their higher stopband performance extends beyond 15 GHz. The designed filters have a flat group delay with less than 0.5 ns peak-to-peak variation across their passbands.

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Figure 6 The measured and simulated S-parameters of the concavely tapered filter. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

Figure 7 The measured group delay of the manufactured filters. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

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NOVEL BROADBAND RECTENNA USING PRINTED MONOPOLE ANTENNA AND HARMONIC-SUPPRESSED STUB FILTER

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ABSTRACT: This report articles a novel broadband rectifying antenna (rectenna) using the printed monopole antenna and the band-pass stub band-pass filter, which is designed in the microstrip line structure at 2.45 GHz for the wireless transmission of microwave power. We suggest a novel H-shaped printed monopole antenna with size reduction of the radiator and the ground plane using surface current distribution, and a broadband stub band-pass filter with suppression of the second harmonic. The rectifying circuit with stub filter doesn’t need additional area because it is fabricated on the ground plane, which is widened to acquire broadband characteristic of the monopole antenna. The RF-to-DC conversion efficiency of 66% is obtained by using a load resistor of 50 Ω. © 2010 Wiley Periodicals, Inc. Microwave Opt Technol Lett 52: 1194–1197, 2010; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.25130

Key words: broadband rectenna; printed monopole antenna; stub BPF; rectifier; wireless power transmission

1. INTRODUCTION

Recently, there is a growing concern on the renewable energy and energy harvesting of industrial infrastructures. The rectifying antenna (rectenna) is one of the important components out of energy harvesting devices and it has been used for providing DC power from space solar power [1]. Also, the rectenna for wireless power transmission has been used in distributed wireless sensor network, RFID tags, and health monitoring of infrastructures [2].

The more antenna size increase, the more DC power can be obtained until the diodes break down; however, because the current portable devices have generally small dimensions, the rectenna should be reduced in size. This leads a small antenna area, and consequently, a low amount of received power. Therefore, the rectenna presented in this article is mainly suitable for low-power applications of wireless power transmission [3].

The typical rectenna in the prior literatures basically consists of four elements: microwave antenna, matching circuit, diode, and DC pass capacitor [1–3]. The initial development of rectenna is focused on antenna gain for more power reception and conversion. The broadband antenna enables relatively high RF power to be received from various sources in the frequency range [3, 4].

This article presents a broadband rectenna, which is designed at 2.45 GHz with the operating range of between 1.8 and 2.8 GHz. Proposed rectenna is combined of a novel printed monopole broadband antenna using surface current distribution and a harmonic-suppressed broadband stub band-pass filter for the suppression the harmonic signals generated from the diode.

2. DESIGN OF A NEW RECTENNA

Usually, the rectenna is consisted of an H-shaped broadband printed monopole antenna, a harmonic-rejecting broadband stub filter for the suppression of the harmonic signals, one detector diode for RF-to-DC conversion, and the DC pass filter. In this article, the proposed rectenna is designed with Ansoft HFSS (ver. 11) and Agilent ADS (ver. 2006). It is fabricated using a standard photolithography process on RT/duriod 6010M substrate, which has a relative dielectric constant of 10.2 and the height of 0.635 mm. Both the antenna feeding and rectifier input port have characteristic impedance of 50 Ω for good impedance matching to reduce signal reflection between these components.

2.1. Antenna Design

The geometry along with its parameters of the novel printed monopole broadband antenna is shown in Figure 1. The antenna has the following dimensions: W = 34 mm, L = 46 mm, Ws = 13.6 mm, Ls = 9 mm, Lc = 3 mm, Wc = 1.6 mm, Wg = 24 mm, and Wf = 1.5 mm. The length of Wg, which can be regarded as a width of dipole antenna is optimized. The more the width is extended, the more broadband characteristic can be achieved. In addition, the rectifying circuit can be implemented on the expanded ground plane. Novel H-shape monopole antenna is compact as compared with the conventional printed monopole antenna. The size reduction can be realized by using surface current distribution as shown in Figure 2. It is based on dipole antenna surface current. Figures 2(a) and 2(b) have the same surface current path length of 57.5 mm, and the size of

Figure 1 Geometry of proposed printed monopole antenna. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
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11. L. Xu, J. Hu, D. Qian, and T. Wang, Novel radio-over-fiber structure with embedded low-pass filter to achieve the required UWB performance [10]. In this article, microstrip-slot-microstrip structure with embedded LPF is utilized to achieve the desired UWB performance [7–13]. A BPF, which utilizes a two-section broadside-coupled microstrip-slotline connected together via a quarter wavelength CPW, achieves a UWB passband [7]. However, the harmonic response is strong at the high cutoff band. In another utilization of broadside-coupled slotline-microstrip structure [8], the slotline in the ground plane is coupled to two microstrip open-circuited stubs on the top layer of the substrate. Because the performance is controlled by the coupling value between the narrow slotline and the microstrip stubs, the device is very sensitive to alignment errors. In another method that relies on the tight coupling of broadside-coupled structures, elliptical shaped microstrip-slot couplers are utilized to construct UWB BPFs [9]. To improve the performance at the high stopband, multiple broadside-coupled sections were utilized. The drawback of this approach is an increased size and a risk of increasing insertion loss. To minimize those adverse effects of the multi section approach, a low-pass filter (LPF) is embedded within the feeder of the filter to improve its high stopband performance without significantly increasing the size, while maintaining the required performance at the passband [10]. In this article, microstrip-slot-microstrip structure with embedded LPF is utilized to achieve the desired UWB performance. This configuration is suitable for the multilayer technology. A complete design method is presented and the final design is tested via simulations and measurements.

2. DESIGN

The configuration of the proposed BPF that utilizes microstrip-slot-microstrip coupled structure is shown in Figure 1. The middle part of the filter is based on the broadside-coupled microstrip-slot-microstrip multilayer configuration. The top layer [Fig. 1(a)] and the bottom layer [Fig. 1(c)] have two similar elliptical shaped microstrip patches that are coupled via an elliptical slot at the middle layer of the structure [Fig. 1(b)]. The ground plane of the whole structure is located in the middle layer. The tapered shaped broadside-coupled patches are utilized as they provide an almost constant tight coupling, which is important to achieve the required filter’s characteristics, across the UWB [15]. When compared with the structure presented in Ref. 9, the main modification in the device shown in Figure 1 is the inclusion of a coupled dumbbell slot in the ground plane [as shown in Fig. 1(b)] and H-shaped shunt open-ended stubs connected with the input and output ports at the top layer as revealed in Figure 1(a). This modification results in an embedded LPF, which improves the high stopband characteristics of the filter if designed properly. The first step in the design procedure of the presented filter is to calculate the required dimensions for the elliptical coupled structure to achieve a passband that extends from 3.1 to 10.6 GHz. To that end, the detailed design method presented in Refs. 9 and 16 is applied. The main parameter needed to calculate the required dimensions is the coupling factor C between the top and bottom layer of the utilized broadside-coupled structure. The structure shown in Figure 1(c) can be considered as a two-section directional coupler connected
using a short microstrip line. Following the quasi-static analysis presented in Ref. 9, it is possible to show that the effective scattering parameters for the two-section broadside-coupled structure ($S_{11ef}$ and $S_{21ef}$) of Figure 1(e) are given as:

$$S_{11ef} = S_{11} + \frac{S_{11}S_{11}}{1 - S_{11}^2}$$  

$$S_{21ef} = \frac{S_{21}^2}{1 - S_{21}^2}$$  

where $S_{11}$ and $S_{21}$ are the return loss and insertion loss, respectively, of each of the two sections, $l_w$ is the physical length of the coupled structure, and $\beta_0$ is the effective phase constant in the medium of the coupled structure. For the configuration under investigation, it is possible to show that

$$\beta_0 = \frac{\beta_e + \beta_o}{2} = \frac{2\pi\lambda}{\sqrt{\epsilon_r}}$$

where $\beta_e$ and $\beta_o$ are the phase constants for the even- and odd-mode respectively, $\lambda$ is the free space wavelength, and $\epsilon_r$ is the dielectric constant of the substrate.

From (1)-(4), it is possible to show that the optimum performance, that is, $S_{11ef} = 0$ and $S_{21ef} = 1$, occurs when the coupling factor between the top and bottom layer ($C$) is equal to 0.707. Using $C = 0.707$ in the quasi-static design approach [9, 16] and assuming that the substrate is Rogers RT6010 (with $\epsilon_r = 10.2$, thickness $= 0.635$ mm, and tangent loss $= 0.0023$) results in the following values for the design parameters depicted in Figure 1; $D_m = 2.8$ mm, $D_s = 5.7$ mm. Concerning length of the coupled structure $l_w$, it is usually chosen to be quarter of the effective wavelength at the center of the passband, that is, 6.85 GHz. Using the software CST Microwave Studio, the optimum value was found to be $l_w = 5.2$ mm. The broadside-coupled structure designed in the previous steps results in a BPF with a wide high stopband. It was found that if length of the slot in the ground plane ($l_s$) and of the shunt stubs at the top layer ($l_{st}$) are equal to 2.5 and 2 mm, respectively, the parameters of the equivalent circuit shown in Figure 2(c) have the following values: $l_s = 1.2$ mm, $C_s = 0.36$ pF, and $R_s = 11$ k$\Omega$. The response of the circuit in this case is as shown in Figure 2(c), which indicates a cutoff band that extends from 11 GHz to more than 30 GHz. The rough design for the embedded LPF is to choose the length of the stub at the top layer and the slot in the ground plane to be half of the effective wavelength at the center of the undesired response. In the broadside-coupled microstrip-slot-microstrip, the harmonic responses

![Figure 1 Configuration of the multilayer broadside-coupled BPF.](image-url)
appears at frequencies, which are twice and three times the frequency of the center passband. Thus, there are two effective spurious responses in the frequency band from 11 to 20 GHz. Therefore, each of the LPFs in Figure 1(e) is designed to effectively remove one of those undesired responses. After optimization using the software CST Microwave Studio, the length of the slot in the ground plane ($l_s$) and of the shunt stubs at the top layer ($l_{st}$) are equal to 2.4 and 2 mm, respectively, for one of the LPFs and 1.6 and 1.4 mm for the other. Width of the slot in the ground plane is fixed at 0.2 mm, whereas the width of the shunt stubs is 0.3 mm, and the radius of the circular slots that form the two ends of the dumbbell slot is 0.3 mm.

3. RESULTS AND DISCUSSION

To verify the performance of the proposed BPF, the device is simulated using the software CST Microwave Studio. A prototype was also manufactured and tested. A photo of the developed filter is depicted in Figure 3. The overall dimension of the designed filter including the microstrip feeders at the input and output is $1.5 \times 2.5$ cm$^2$. Because of the use of coupled patches at the top and bottom layers, there is a possibility of limited radiation from the filter. This radiation can have adverse effects on the performance, such as increasing the insertion loss or level of the harmonic responses. To prevent those adverse effects, the filter was simulated and measured while it was enclosed in a shielding box of dimensions $1.5 \times 2.5 \times 1$ cm$^3$. A comparison is also made with the performance without the enclosure. Variation of the simulated and measured scattering parameters ($S_{11}$ and $S_{21}$) with frequency is shown in Figure 4. It is apparent that the designed filter has a passband that covers the UWB range of 3.1–10.6 GHz assuming the 3-dB insertion loss as a reference. The insertion loss at the center of the passband is less than 0.3 dB, according to the simulations and less than 0.5 dB in the measured results, whereas the return loss is larger than 15 dB.

Also, the results of Figure 4 reveal a sharp cutoff at the upper stopband between 10.6 and 14 GHz indicating the effectiveness of the embedded LPFs. In general, Figure 4 reveals a good agreement between the simulated and measured results. To prove the benefits of shielding the device, the filter was also simulated in case there is no enclosure. The results concerning the insertion loss is included in Figure 4. It is clear that without the enclosure, the filter has an additional 0.2-dB insertion loss when compared with the simulation results with enclosure. Moreover, three residual harmonic responses appear at around 13.7, 16.2, and 19.3 GHz. One of the important parameters that define the quality of the performance of filters is the variation in the group delay across their passband. For impulse radio systems, the filter is required to have a flat group delay within its passband. For the BPF presented in this article, the group delay was measured, and the result is depicted in Figure 4. The device has about 0.3-ns peak-to-peak variation in the group delay across the band 3–10 GHz indicating a low time domain distortion. Finally, if the performance of the two-section broadside-coupled BPF with embedded LPFs presented in this article is compared with that of the two-section BPF without the embedded structure [9], it will become obvious that despite the fact that the two devices have the same size, the modified structure has a sharper and wider high cutoff band.

Figure 2 Configuration of the LPF (a), its equivalent circuit (b), and performance (c). [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 3 Photo of the manufactured filter. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 4 The simulated and measured performance of the filter. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
4. CONCLUSIONS

The design of a BPF using a multilayer broadside-coupled structure has been presented. The UWB performance of the filter is achieved via the use of controlled coupling between microstrip elliptical patches at the top and bottom layers through a slot at the middle layer. To widen the high cutoff band of the filter, H-shaped shunt open-ended connected with the microstrip line of the top layer and coupled with dumbbell shaped slots at the ground plane of the middle layer are used to form two embedded LPFs. Those LPFs are designed to remove any harmonic responses that may appear beyond 10.6 GHz. The simulated and measured results of the filter prove its UWB passband and a high cutoff band that extends beyond 20 GHz. The designed filter also has a flat group delay with less than 0.3-ns peak-to-peak variation across its passband.

REFERENCES


A CMOS QPSK DEMODULATOR FRONTEND FOR A PON ONU

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ABSTRACT: A quadrature phase-shift keying (QPSK) demodulator frontend for a passive optical network (PON) optical network unit receiver fabricated in CMOS 0.13-μm technology is presented. The frontend integrates a compact 5-GHz ring quadrature voltage-controlled oscillator (QVCO), two inductor-less broadband mixers, and two second-order LC ladder low-pass filters. The broadband mixers operate with an input frequency between 2.5 and 7.5 GHz and with an input 1-dB compression point of ~5 dBm. Area-saving design techniques are applied to lower the cost of the terminal receivers, because a competitive cost is a key requirement for the deployment of the PON access network. The frontend circuit presented achieves an area savings of more than 90% for the QVCO and mixers when compared with some published designs that can also fit the application. Measurement and simulation results are presented to verify that this frontend can be used to demodulate a QPSK signal with a data rate as high as 5 Gb/s, which is twice the downstream data rate of the current Gigabit PON standard. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 53:1056–1062, 2011; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.25902

Key words: CMOS; PON; QPSK; demodulator; frontend

1. INTRODUCTION

Passive optical networks (PON) are methods promising implementation for broadband access networks like fiber to the home/premises (FTTH/P). PONs require no electrical power supply between the central office, optical line terminal, and the end users, optical network units (ONUs). The ITU-T G.984 Gigabit PON (GPON) standard [1] has defined an industry best practice with a downstream bandwidth of 2.488 Gb/s and an upstream bandwidth of 1.244 Gb/s. PON ONU transceivers are very cost sensitive because the end users will directly bear the cost of ONU equipment located in their home or premises. An integrated ONU solution has been proposed in Ref. 2, which uses a novel offset-sidewave modulated optical signal. This technique is used to create a quadrature phase-shift keying (QPSK) signal. The integrated ONU consists of a photonic integrated circuit functioning as an optical-electrical interface and a microelectronic integrated circuit (MIC) functioning as electrical transceiver. The receiver in the MIC consists of a bandpass transimpedance amplifier (TIA) and a direct down-conversion demodulator.

This article proposes a demodulator frontend for the proposed receiver in Ref. 2. It consists of two mixers, a quadrature voltage-controlled oscillator (QVCO), and two low-pass filters. This frontend can be used in a four-phase Costas loop [3] as shown in Figure 1, which can be used for the simultaneous carrier recovery and demodulation of a QPSK signal. In addition to the frontend circuit discussed in this article, a Costas loop contains two limiting amplifiers, two baseband multipliers, one subtractor, and one loop filter.

2. COMPONENT DESIGN

This section describes the circuit design and measurement results for the QVCO and the broadband mixers separately.

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ABSTRACT: A quadrature phase-shift keying (QPSK) demodulator frontend for a passive optical network (PON) optical network unit receiver fabricated in CMOS 0.13-μm technology is presented. The frontend integrates a compact 5-GHz ring quadrature voltage-controlled oscillator (QVCO), two inductor-less broadband mixers, and two second-order LC ladder low-pass filters. The broadband mixers operate with an input frequency between 2.5 and 7.5 GHz and with an input 1-dB compression point of ~5 dBm. Area-saving design techniques are applied to lower the cost of the terminal receivers, because a competitive cost is a key requirement for the deployment of the PON access network. The frontend circuit presented achieves an area savings of more than 90% for the QVCO and mixers when compared with some published designs that can also fit the application. Measurement and simulation results are presented to verify that this frontend can be used to demodulate a QPSK signal with a data rate as high as 5 Gb/s, which is twice the downstream data rate of the current Gigabit PON standard. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 53:1056–1062, 2011; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.25902

Key words: CMOS; PON; QPSK; demodulator; frontend

1. INTRODUCTION

Passive optical networks (PON) are methods promising implementation for broadband access networks like fiber to the home/premises (FTTH/P). PONs require no electrical power supply between the central office, optical line terminal, and the end users, optical network units (ONUs). The ITU-T G.984 Gigabit PON (GPON) standard [1] has defined an industry best practice with a downstream bandwidth of 2.488 Gb/s and an upstream bandwidth of 1.244 Gb/s. PON ONU transceivers are very cost sensitive because the end users will directly bear the cost of ONU equipment located in their home or premises. An integrated ONU solution has been proposed in Ref. 2, which uses a novel offset-sideband modulated optical signal. This technique is used to create a quadrature phase-shift keying (QPSK) signal. The integrated ONU consists of a photonic integrated circuit functioning as an optical-electrical interface and a microelectronic integrated circuit (MIC) functioning as electrical transceiver. The receiver in the MIC consists of a bandpass transimpedance amplifier (TIA) and a direct down-conversion demodulator.

This article proposes a demodulator frontend for the proposed receiver in Ref. 2. It consists of two mixers, a quadrature voltage-controlled oscillator (QVCO), and two low-pass filters. This frontend can be used in a four-phase Costas loop [3] as shown in Figure 1, which can be used for the simultaneous carrier recovery and demodulation of a QPSK signal. In addition to the frontend circuit discussed in this article, a Costas loop contains two limiting amplifiers, two baseband multipliers, one subtractor, and one loop filter.

2. COMPONENT DESIGN

This section describes the circuit design and measurement results for the QVCO and the broadband mixers separately.
Ultra-wideband bandpass filters using broadside-coupled microstrip–coplanar waveguide

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Abstract: Bandpass filters that cover the ultra-wideband (UWB) frequency range (3.1–10.6 GHz) are presented. The filters utilise broadside-coupled microstrip–coplanar waveguide making them suitable for printed circuit board technology. To achieve a wide upper stopband, radial slots and stepped impedance resonators are employed either to suppress or to relocate the harmonic responses outside the band of interest. The presented design procedure relies on a quasi-static analysis and conformal mapping. The simulated and measured results of the proposed filters prove their suitability for UWB. Furthermore, the presented filters have a wide upper stopband that extends above 20 GHz, a compact size and less than 0.12 ns peak-to-peak variation in the group delay across the passband.

1 Introduction

In 2002, the Federal Communications Commission (FCC) assigned the frequency band of 3.1–10.6 GHz to emerging ultra-wideband (UWB) technologies and their applications. The primary objective of UWB is the possibility of achieving high data rate communication in the presence of existing wireless communication standards. To meet the FCC spectrum mask, UWB sub-systems require bandpass filters (BPFs) which feature low insertion loss over the passband (3.1–10.6 GHz), a flat group delay performance within that band and a good rejection characteristic outside the band. To meet these requirements, different types of UWB BPFs have been reported in the literature [1–24].

One of the main classes of BPFs relies on parallel or edge-coupled structures [1–7]. However, the tolerance of the microstrip and coplanar waveguide (CPW) fabrication process imposes an upper limit upon the coupling levels for parallel- and edge-coupled structures. This makes the manufacturing process for the UWB filters utilising those structures difficult as their performance is very sensitive to manufacturing errors. This difficulty can be overcome by utilising broadside-coupled structures that can offer tight coupling across UWB [8–20].

In [8], a UWB BPF was presented by utilising a broadside-coupled microstrip–CPW transition, which was proposed previously [9]. However, the results show that up to three sections should be connected together to achieve the required passband performance. Moreover, the filter suffers from a serious harmonic response at the high out-of-band frequencies. The authors had to add another section to the three-section filter to cancel the effects of the harmonic responses and to extend the upper stopband. A close look at the design presented in [8] shows that it does not rely only on the broadside coupling between a microstrip line and a CPW, but it also depends on the edge coupling as the gap between the input and output ports is only 0.2 mm.

In another design, which also relies on a microstrip–CPW transition, a hybrid microstrip and CPW structure that utilises a multiple-mode resonator was proposed to design a UWB BPF [10]. The device consists of three sections; two broadside-coupled microstrip–CPW structures and a half-wavelength multiple-mode resonator. The total length of these three sections is equal to, or more than, one wavelength, which makes the filter relatively large in size. From the results at around 13 GHz, it is apparent that the device has a strong harmonic response at and above 13 GHz which limits the high out-of-band rejection. A similar design with similar drawbacks is also presented in [11].

A more compact UWB BPF design utilising a two-section broadside-coupled microstrip–slotline connected together via a quarter wavelength CPW is described in [12]. However, the harmonic response is strong at the high cutoff band. In another utilisation of broadside-coupled slotline-microstrip structure [13], the slotline in the ground plane is coupled to two microstrip open-circuited stubs on the top layer of the substrate. Since the performance is controlled by the coupling value between the narrow slotline and the microstrip stubs, the device is very sensitive to alignment errors. The presented results are given only for limited high out-of-band frequencies. Thus, it is not possible to verify whether or not the filter has harmonic responses at a wide high band.

In another class of UWB BPFs, multilayer technology is employed to achieve the required tight broadside coupling. In [14], elliptical-shaped broadside microstrip–slot couplers [15] are used to construct UWB BPFs. To improve the performance at the high stopband, multiple broadside-coupled sections of up to five were utilised. The drawback of using five sections is an increased size and a noticeable increase in the insertion loss at the passband. In a recent
This design requires the use of sophisticated tools as the distances between different parts of the structure can be as low as 0.02 mm. Moreover, the results show high insertion loss (2 dB) at part of the passband (4 GHz). In another approach to the design of liquid crystal polymer-based BPF [18], broadside-coupled microstrip radial stubs and high-impedance microstrip lines on three-metal-layer structures are utilised to realise the required UWB performance. To improve the performance at the upper stopband, several stubs with short-circuit vias are needed.

The use of another technology, namely low-temperature co-fired ceramics (LTCC), for the design of different types of BPFs is investigated in [19, 20]. The procedure presented in [19], which uses multiple transmission lines shunted with short-circuited stubs, resulted in two relatively narrow bandpass regions, which does not suit UWB applications. The three-section stripline-coupled structure utilised in LTCC [20] shows an UWB performance. The offset distance between the top and bottom layer, which is as low as 0.02 mm, is used to control the value of the coupling, and thus, the performance of the filter. This very small spacing puts a huge burden on the manufacturing process, as it requires costly tools.

In another development, a UWB BPF that relies on broadside-coupled CPW–CPW and stripline–stripline structures is presented in [21]. The filter needs several stubs and bonding wires to achieve the required performance. The proposed configuration presents several manufacturing challenges as the value of the coupling for the presented filter is very sensitive to the offset distance between the top and bottom strips, having widths as low as 0.2 mm.

Different combinations of Y-shaped resonators in the edge-coupled and broadside-coupled topologies have recently been used to build UWB BPFs [22]. The main drawback of the edge-coupled topology is the narrow gap needed for a tight coupling. Moreover, the presented results show a strong harmonic response at 18 GHz.

The above considerations of the existing designs of UWB BPFs indicate that most of them use a multilayer broadside coupling approach to meet the requirement for a tight coupling. Their frequent shortcomings include very small spacing of the layers in some of the utilised techniques leading to costly manufacturing, the use of a large number of sections resulting in increased insertion losses and poor out-of-band rejection characteristics accompanied by the presence of spurious harmonics.

This paper addresses the above problems by reporting the design of a tapered broadside-coupled microstrip–CPW UWB BPF structure as an alternative to the multilayer configurations. The proposed configuration suits the use of simple printed circuit board (PCB) technology. Thus, it alleviates the need to use the expensive multilayer technology and its sophisticated fabrication process. In the filters presented here, embedded radial slots and stepped impedance resonator (SIR) are employed to achieve a wide upper stopband. A complete design method is presented for the proposed devices and the final designs are tested via simulations and measurements.

2 Theory and design

The configuration of the BPF that utilises a microstrip–CPW coupled structure is shown in Fig. 1. Tapered coupled lines of elliptical shape are used because they provide an almost constant tight coupling across a large frequency range [25], which is important to achieve a flat bandpass characteristic of filter over the UWB. The proposed configuration shows some similarities with the multilayer microstrip–slot–microstrip approach [14, 16] whose drawbacks, such as challenges of aligning the different layers and the accompanying costs, are avoided. In the new approach, the coupling patch in the bottom layer is included within the slot of the mid-layer. Thus, the PCB of the bottom layer, that is, the third layer in the multilayer structure of [14], is removed resulting in a simple two-sided PCB. To minimise insertion losses, present in the multi-section designs reviewed earlier, the filter uses a two-section broadside-coupled microstrip–CPW structure. Those two sections are connected together via a short length of CPW in the bottom layer.

In the following, we present the analysis for rectangular-shaped coupled sections, whereas the real elements used in the filter’s structure are elliptically shaped (Fig. 1). This is required for the ease of derivations via conformal mapping. The final results for the elliptically-shaped coupled lines can be obtained using equivalence between the elliptical and rectangular-shaped line sections [15].

Following the analysis presented in [14], it is possible to show that the effective scattering parameters for the two-section broadside-coupled structure (S11ef and S21ef) shown
in Fig. 1c are given as

\[
S_{11}^{ef} = S_{11} + \frac{S_{12}^2 S_{11}}{1 - S_{11}}, \quad S_{21}^{ef} = \frac{S_{21}^2}{1 - S_{11}} \tag{1}
\]

\[
S_{11} = \frac{1 - CF^2 (1 + \sin^2 (\beta_{ef} l))}{\sqrt{1 - CF^2 \cos (\beta_{ef} l) + j \sin (\beta_{ef} l)}} \tag{2}
\]

\[
S_{21} = \frac{j2CF \sqrt{1 - CF^2 \sin (\beta_{ef} l)}}{\sqrt{1 - CF^2 \cos (\beta_{ef} l) + j \sin (\beta_{ef} l)}} \tag{3}
\]

where \(S_{11}\) and \(S_{21}\) are the scattering parameters for one section of the coupled structure, \(CF\) is the coupling factor between the top layer and the bottom layer of a one-section broadside-coupled structure, \(\beta_{ef}\) is the effective phase constant in the medium of the coupled structure, and \(l\) is the physical length of the coupled structure which is chosen such that \(\beta_{ef} l = \pi/2\) at the centre of the passband (6.85 GHz).

It is possible to solve (1)–(3) to find that \(CF = 1/\sqrt{2}\) gives the best performance, that is, \(S_{11}^{ef} = 0\) and \(S_{21}^{ef} = 1\), at the centre of the passband of the device. Thus, this value for \(CF\) is used in the design of the filters presented in this paper.

To find the relation between the coupling factor \(CF\) and the physical dimensions of the filter, each of the two sections shown in Fig. 1 can be analysed using the odd- and even-mode approach. The excitations needed to generate the two modes and distribution of lines of the electric field between the two broadside-coupled layers for the two fundamental modes is shown in Fig. 2.

For the even-mode, the two layers are excited in-phase, whereas in the odd-mode, the top and bottom layers are out-of-phase with respect to the ground plane. Assuming a quasi-transverse electromagnetic propagation, the electrical characteristics of the coupled lines can be completely determined from the effective per unit length capacitances of the lines and the phase velocity on the lines [26]. Therefore equivalent circuits shown in Fig. 3 can be used to analyse the proposed filter.

For each of the two modes of propagation, the capacitance for each of the two coupled lines can be determined from Fig. 3. The even-mode capacitance for the microstrip \((C_{me})\) and the CPW \((C_{ce})\) are equal to

\[
C_{me} = C_{mg}, \quad C_{ce} = C_{eg} \tag{4}
\]

The odd-mode capacitance for the microstrip \((C_{mo})\) and the CPW \((C_{co})\) are equal to

\[
C_{mo} = C_{mg} + 2C_{mc}, \quad C_{co} = C_{eg} + 2C_{mc} \tag{5}
\]

The characteristic impedance of each of the two lines (microstrip at the top layer and CPW at the bottom layer) at any of the two modes can be found as follows [27]

\[
Z_{ij} = 1/(\nu C_{ij}) \tag{6}
\]

where the subscript \(i\) refers to the line (\(m\) for microstrip and \(c\) for CPW) and \(j\) refers to the mode (\(e\) for even and \(o\) for odd), \(\nu\) is the phase velocity \((\nu = c/\sqrt{\varepsilon_r})\) and it is assumed to be the same for the two modes for the structure under consideration.

The structure shown in Figs. 1–3 is asymmetrical. Therefore the analysis method for asymmetrical broadside-coupled lines [27] is used. According to that method, the coupling factor between the top layer and the bottom layer \((CF)\) can be found as

\[
CF = \frac{Z_{mc} Z_{mo}}{Z_{ce} Z_{co} \sqrt{(Z_{mc} + Z_{mo})(Z_{ce} + Z_{co})}} \tag{7}
\]

It is possible to use a similar analysis to the one in [26], to show that in order to achieve a perfect matching at the input/output ports, which have characteristic impedance \(Z_0(=50\, \Omega)\), then

\[
Z_0 = \sqrt{Z_{mc} Z_{mo}} \tag{8}
\]

Substituting (8) into (7) gives the following result

\[
CF = \frac{Z_0 (Z_{ce} - Z_{co})}{\sqrt{(Z_{ce} Z_{co})(Z_{mc} + Z_{mo})(Z_{ce} + Z_{co})}} \tag{9}
\]

The coupling factors as a function of the capacitances can be obtained by substituting (6) into (9).
The dimensions of the coupled microstrip–CPW offer the required capacitances, as from (7) to (9), and thus coupling factors can be determined by utilising the quasi-static approach. Using that approach with the help of the conformal mapping technique [28, 29], the capacitances shown in Fig. 3 can be calculated as a function of the coupled structure’s dimensions. The final equations for the capacitances per unit length are

\[ C_{mg} = 2\varepsilon_0\varepsilon_r K'(k_1) \]  
\[ C_{cg} = 2\varepsilon_0(1 - \frac{K(k_2)}{K'(k_2)} + \frac{K(k_1)}{K'(k_1)}) \]  
\[ C_{mc} = \varepsilon_0\varepsilon_r \frac{D_m + D_c}{2h} \]  
\[ k_1 = \frac{\sinh(\pi D_c/(4h))}{\sqrt{\sinh^2(\pi D_c/(4h)) + \cosh^2(\pi D_m/(4h))}} \]  
\[ k_2 = \frac{\sinh(\pi D_c/(4h))}{\sinh(\pi D_c/(4h))} \]  
\[ k_3 = D_c/D_s \]  

where \( K(k) \) is the first kind elliptical integral and \( K'(k) = K(\sqrt{1 - k^2}) \), \( D_m \), \( D_c \) and \( D_s \) are the diameters shown in Fig. 1 and \( h \) is the thickness of the substrate.

The design (6)–(15) can now be used to find the initial dimensions \( (D_m, D_c \) and \( D_s \) of the BPF assuming \( CF = 1/\sqrt{2} \). All of the solutions obtained using (6)–(15) are guaranteed to give a theoretically ideal performance at the centre of the passband. However, there is no guarantee that all these solutions provide the required performance across the whole band under consideration, say from DC up to 20 GHz. Thus, the available solutions from (6) to (15) are checked using a full-wave electromagnetic simulation to see which one of them achieves the best performance across the band under investigation.

Concerning the other physical dimensions of the filter, the length of the coupling slot in the ground plane \( (l_3) \) is chosen according to the design principles of coupled structures to be equal to the quarter of the effective wavelength at the centre of the passband \( (6.85 \text{ GHz}) \), whereas the lengths \( (l_1 \) and \( l_2) \) are chosen to be less than \( l_3 \) by the value of the slot needed in the ground plane to achieve a perfect matching between the coupled structure and the input/output ports. The length of the CPW connecting the two coupled structures at the ground plane \( (d) \) can be used to fine adjust width of the passband [14].

Using the above derivations, the initial physical dimensions of the filter are worked assuming Rogers RT6010 substrate with \( \varepsilon_r = 10.2 \) and thickness = 0.635 mm. Next, the software computer simulation technology (CST) Microwave Studio is used to further tune the dimensions, which are equal to \( D_m = 2 \text{ mm}, \ D_c = 4.2 \text{ mm}, \ D_s = 5.3 \text{ mm}, \ l_1 = 4 \text{ mm}, \ l_2 = 4.1 \text{ mm}, \ l_3 = 5.7 \text{ mm} \) and \( d = 3.5 \text{ mm} \).

There is an additional step in the design procedure needed to ensure the complete cancellation of the harmonic responses at a wide upper stopband. Two methods are considered to achieve that goal. In the first approach, which is used traditionally, a low-pass filter in the form of radial slots is inserted in the ground plane of the CPW line connecting the two sections as depicted in Fig. 4.

It is shown [30] that quarter-wavelength, shunt radial stubs in microstrip line or radial slots (in CPW) behave as a wide bandstop filter. The radial slot is usually designed to cause an attenuation pole at the mid-band frequency of the stopband. The attenuation-pole frequency is decided by the slot’s radius, whereas the attenuation bandwidth is defined by the radial angle. A larger radius results in a lower attenuation-pole frequency, whereas a larger radial angle causes a reduction in the reactance slope of the slot’s input reactance, and thus an increase in the bandwidth of attenuation [30].

In this paper, the initial values for the radii of the utilised radial slots are chosen to be around quarter of a wavelength calculated at the centre of the undesired responses. Those initial values are then optimised to obtain the best possible rejection of the harmonic response without changing the performance at the passband. The optimised dimensions for the filter, in this case using RT6010 as a substrate, are \( D_m = 2 \text{ mm}, \ D_c = 2.2 \text{ mm}, \ D_s = 4.1 \text{ mm}, \ l_1 = 2.5 \text{ mm}, \ l_2 = 2.8 \text{ mm}, \ l_3 = 5.5 \text{ mm} \) and \( d = 3.5 \text{ mm} \). The radii of the 60˚ radial slots are between 0.9 and 1.6 mm.

The second approach to remove the harmonic responses is to convert the CPW connecting the two sections into a SIR. Thus, the two sections that form the BPF are overlapped in the manner shown in Fig. 5. This configuration can be represented by a two-section coupled structure connected using a SIR. The SIR can be designed specifically to cancel the effects of the spurious responses [31]. Let us first assume that \( Z_e = \) the ratio of the impedance of the two ends of the SIR to the impedance of the middle section of the SIR. The required length of the SIR needed to resonate at a
frequency that has an effective λs, which is equal to [31]

\[ l_{\text{SIR}} = 2\lambda_s \tan^{-1} \sqrt{Z_r / \pi} \]  \hspace{1cm} (16)

It is clear from (16) that if a uniform resonator is used, that is, \( Z_r = 1 \), the length of the resonator is \( \lambda_s / 2 \). However, if \( Z_r < 1 \), \( L_{\text{SIR}} < \lambda_s / 2 \), and thus a compact resonator can be used. The other important parameter in the design of the SIR is the location of the harmonic responses. From [31], values of the first three harmonic responses \( (f_{h1}, f_{h2}, f_{h3}) \) relative to the resonant frequency \( f_o \) are

\[ f_{h1} / f_o = \pi / (2 \tan^{-1} \sqrt{Z_r}) \] \hspace{1cm} (17a)

\[ f_{h2} / f_o = 2(f_{h1} / f_o) - 1 \] \hspace{1cm} (17b)

\[ f_{h3} / f_o = 2(f_{h1} / f_o) \] \hspace{1cm} (17c)

The target of the design is to have \( f_o \) at the centre of the passband, that is, at 6.85 GHz, and each of the harmonic frequencies \( (f_{h1}, f_{h2}, f_{h3}, \ldots) \) to be larger than 20 GHz. Solving (17) results in \( Z_r \) to be 0.32. If \( Z_r \) is taken as 0.3 for an additional safety margin, the harmonic responses appear at \( f_{h1} = 21.5 \) GHz, \( f_{h2} = 36 \) GHz and \( f_{h3} = 42.85 \) GHz. The total length of the SIR needed to achieve those characteristics is 0.32λ according to (16). With this SIR, the upper stopband extends up to 21.5 GHz. For the applications that require a wider upper stopband, \( Z_r \) is to be chosen well below 0.3.

For the overlapped structure of Fig. 5, the two coupled parts connected at the ends of the SIR are designed following the procedure of the broadband-coupled microstrip–CPW (6)–(15). After obtaining the initial dimensions, the overall structure is tuned for an optimum performance using CST Microwave Studio. The final dimensions of the filter after manual tuning are \( D_m = 1.8 \) mm, \( D_c = 3.8 \) mm, \( D_s = 4.6 \) mm, \( l_1 = 1.5 \) mm, \( l_2 = 3.1 \) mm, \( l_3 = 7 \) mm and \( l_c = 12.2 \) mm. It was noticed that the tuned dimensions were different by around ±10% compared with the initial values. For example, the initial values for \( D_m, D_c \) and \( D_s \), calculated using (6)–(15) are 1.74, 3.94 and 4.13 mm, respectively. By using the overlapped structure, the device has a compact size with an overall dimension including the input/output feeders of 1.2 cm × 1.8 cm.

### 3 Results and discussions

The designed filters were tested via simulations using CST Microwave Studio. A prototype of each of the designed filters was also developed using the substrate RT6010 and an automated milling machine with ±0.1 mm accuracy. A photograph of one of the developed devices is depicted in Fig. 6. The simulated and measured results for the BPF designed using the broadband-coupled microstrip–CPW structure and embedded radial slots are shown in Fig. 7. Note that the measured results include the effect of the subminiature A connectors used during the testing. The results in Fig. 7 reveal a passband from 3.1 to 10.6 GHz. The insertion loss is less than 0.5 dB, whereas the return loss is larger than 15 dB, across most of the UWB passband. The filter has a wide upper stopband that extends above 20 GHz. Fig. 6 also shows a sharp cutoff at the frequency band 11–14 GHz. During the tests, the device was enclosed in a shielding box with dimensions of 1.5 cm × 2 cm × 1.5 cm to prevent any radiation losses. When the filter was tested without using the shielding box, the measured insertion loss of the filter was around 0.8 dB across most of the passband. It is to be noted from Fig. 7 that there is, in general, a good agreement between the simulated and measured results.

For the impulse radio systems, the BPF should have a flat group delay across the passband to keep the distortion of the pulse shape to minimum. Thus, variation of the group delay of the designed filter should have a sub-nanosecond peak-to-peak variation across the passband. The measured results of the group delay for the manufactured filter are shown in Fig. 7. There is a 0.12 ns peak-to-peak variation in the group delay across the UWB passband. Concerning the overlapped structure of Fig. 5, this was tested while the device was enclosed in a shielding box of dimensions (1.2 cm × 1.8 cm × 1 cm) to minimise effects of any radiation from the coupled patches. The results in Fig. 8 indicate a passband that covers the range 3.1–10.6 GHz assuming the 3 dB insertion loss as a reference. The insertion loss is less than 0.5 dB and the return loss is larger than 15 dB across most of the passband. Moreover, the filter has a wide stopband that extends beyond 20 GHz. The harmonic responses are removed entirely from the band of interest, that is, up to 20 GHz. This result proves accuracy of the design procedure adopted for this structure. The measured results depicted in Fig. 8 confirm the simulated performance, although there is an additional 0.3 dB insertion loss at the passband.

To confirm the validity of the presented design method, the performance of the overlapped filter designed using the initial calculated values is included in Fig. 8. It is clear that those initial values result in a wideband BPF that has a passband extending from 3.2 to 9.4 GHz. From the results, it is apparent that the manual optimisation involving an iterative use of CST Microwave Studio gives an improved passband that covers the frequency range from 3.1 to 10.6 GHz.
Fig. 8 Performance of the overlapped bandpass filter

The other important parameter needed to judge performance of the overlapped BPF of Fig. 5 is the group delay. The measured results of the group delay for the manufactured filter depicted in Fig. 8 reveal 0.12 ns peak-to-peak variation in the group delay across the UWB passband.

From a comparison between the performance of the BPF with embedded radial slots (Fig. 7) and that of the overlapped structure (Fig. 8), it is apparent that the two topologies offer almost the same performance at the lower stopband and the passband. However, there is a significant difference in the performance at the upper stopband. For the BPF with embedded radial slots, the cutoff is sharper near the band 13 GHz, but there are residuals of the harmonic responses, although they are very low in amplitude, near 15 and 18 GHz. For the overlapped structure, the harmonics are removed entirely from the upper stopband under investigation. This is in an exact agreement with the utilised design procedure that aimed at relocating the harmonic response to above 20 GHz.

4 Conclusion

BPFs that cover the UWB frequency (3.1–10.6 GHz) have been presented. These filters are based on a broadside-coupled microstrip patch at the top layer of a substrate and a CPW at the bottom layer. Thus, the proposed filters can be manufactured easily by using PCB technology. To remove the harmonic responses from the upper stopband, and thus to achieve a wide upper stopband, two methods have been utilised; in the first one, radial slots with low-pass characteristics are embedded in the CPW, whereas, in the second method, two sections of a broadside-coupled microstrip–CPW BPF are overlapped to form an embedded SIR between them. The SIR is employed to relocate the harmonic responses outside the band of interest. A detailed design procedure has been presented for the proposed filters. The simulated and measured results have proven the UWB performance of the presented devices with a wide upper stopband that extends above 20 GHz. These filters have less than 0.12 ns peak-to-peak deviation in the group delay across the passband enabling their use in the pulsed radio systems with distortionless operation. Their small size enables their use in compact UWB sub-systems.

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Ultrawideband Balanced Bandpass Filter
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Abstract—A broadside-coupled structure is utilized to design balanced bandpass filter (BPF) with ultrawideband performance. The top and bottom layers of the structure contain tapered microstrip patches. Those patches are coupled via tapered slots in the ground plane, which is located at the middle layer. The employed structure operates as a BPF in the differential-mode, whereas it operates as an all-stop filter in the common-mode. The simulated and measured performance of the filter show a 123% fractional bandwidth centered at 6.55 GHz, a sharp and wide upper stopband that extends beyond 20 GHz, and a sharp lower stopband. The filter suppresses the common-mode signals by more than 27 dB in the simulations and 24 dB in the measurements across the whole abovementioned band. The designed filter reveals less than 27 dB in the simulations and 24 dB in the measurements of the attenuation. The presented filter has a flat group delay across its passband. The performance of the presented filter is confirmed via simulations and measurements.

Index Terms—Balanced filter, bandpass filter (BPF), differential filter, ultrawideband (UWB).

I. INTRODUCTION

Balanced filters are essential in building communication systems due to their crucial role in reducing the interference, noise and crosstalk between different elements of the system. The modern trend in building filters, amplifiers, mixers, and oscillators is to use the balanced configuration. The balanced transceivers that include those devices have higher immunity to noise and interference and a higher signal-to-noise ratio in comparison to the single-ended transceivers.

The balanced bandpass filters (BPF) with differential input/output ports should have the required passband and stopbands in the differential-mode operation, and be effective in removing the common-mode signals across at least their passbands. Those characteristics are essential to ensure a high signal-to-noise ratio in communication systems.

Due to the enormous increase in demand for wideband applications, such as the ultrawideband (UWB) 3.1 to 10.6 GHz systems, the design of balanced filters with wideband or multi-band performance has attracted a significant attention recently [2]–[7].

Different configurations, such as coupled stepped-impedance resonators [1], end- and broadside-coupled coplanar striplines [2], branch-line structures [3], double-sided parallel-strip lines [4], vertical split ring resonators based on low-temperature co-fired ceramics [5], and metamaterials [6], were utilized to design balanced filters. The measured and simulated performance of the presented balanced filters shows that they have either a limited bandwidth that does not suit the UWB applications [1], [3], [6], a narrow upper stopband [4], or a high insertion loss across most of the passband [2], [5].

In this letter, a broadside-coupled microstrip-slot-microstrip structure is utilized to develop an UWB balanced BPF. The filter is designed to have a passband in the differential-mode of more than 120% fractional bandwidth. In the common-mode, the proposed filter behaves as an all-stop filter with more than 24 dB of attenuation across the band (3.1 to 10.6 GHz). The lower and upper cutoff bands show sharp rejection characteristics, with upper stopband extending to 20 GHz. Moreover, the presented filter has a flat group delay across its passband. The performance of the proposed filter is confirmed via simulations and measurements.

II. THE PROPOSED DEVICE

The multilayer broadside-coupled structure depicted in Fig. 1 is employed to build an UWB balanced filter. In the utilized structure, two pairs of similar tapered microstrip patches at the top and bottom layers are coupled through a pair of tapered slots at the middle layer of the structure, which also contains the ground plane. There are two vias connecting the far ends of the coupled microstrip patches to the ground plane as shown in Fig. 1. The utilized structure includes three conductive layers interleaved by two substrates.

It has been shown that a BPF based on a broadside-coupled structure can be easily designed to produce a wide passband. However, it has a narrow upper stopband with a slow cutoff rate [7]. Those conclusions are applied to the proposed balanced filter that uses a broadside-coupled structure. Thus, to improve the sharpness of the stopbands and to make the filter’s ports uniplanar, two sections are employed as revealed in Fig. 1. To widen the upper stopband by removing the harmonic responses, a lowpass filter (LPF) is embedded in the microstrip line connecting the two utilized sections of the filter as shown in Fig. 1. H-shaped shunt open-ended stubs connected with the microstrip line of the bottom layer and coupled with dumbbell shaped slots at the ground plane of the middle layer are employed to form the LPF [8].

To explain the fundamental of operation for the proposed filter, a simplified diagram for the device is deduced from Fig. 1(a) and depicted in Fig. 1(e). In the differential-mode operation, a virtual short circuit appears along the axis of symmetry indicated in Fig. 1(a) and (e). In this case, section #1 and section #2 are connected virtually to the ground at points $p_1$ and $p_2$, respectively. Using the four-port network analysis...
and transmission \((T)\) coefficients for each of the coupled sections in this mode are

\[
\Gamma = \frac{1 - \zeta^2 (1 + \sin^2(\xi_f I^2_z))}{\sqrt[1]{1 - \zeta^2 \cos(\xi_f I^2_z) + j \sin(\xi_f I^2_z)}}^2 \tag{1}
\]

\[
T = \frac{j \zeta^2 \sqrt[1]{1 - \zeta^2 \sin(\xi_f I^2_z) + j \cos(\xi_f I^2_z)}}{\sqrt[1]{1 - \zeta^2 \cos(\xi_f I^2_z) + j \sin(\xi_f I^2_z)}}^2 \tag{2}
\]

\(\zeta\) is the coupling factor between the top and bottom layers, \(\xi_f\) is the effective phase constant in the medium of the coupled structure, and \(I^2_z\) is the physical length of the coupled structure [Fig. 1(a)], which is equal to quarter of the effective wavelength calculated at the center of the passband.

In the common-mode operation, a virtual open circuit appears along the axis of symmetry, and thus, at \(p_1\) and \(p_2\) marked in Fig. 1(e), \(\Gamma\) and \(T\) for each of the coupled sections in this mode can be found using the four-port network analysis

\[
\Gamma = 1; \ T = 1. \tag{3}
\]

Thus, the structure in the common-mode behaves as an all-stop filter. This is the preferred option for the behavior of the balanced circuit in the common-mode when comparing with having a bandstop filter. This is because the all-stop filter secures the rejection of the common-mode signals across the whole passband of the BPF, whereas the bandstop filter usually has a low rejection capability at the lower and upper ends of its stopband.

The design procedure for the proposed filter starts by finding the suitable value for the coupling factor \(\zeta\). Solving (1)–(2) shows that the optimum performance \((\Gamma = 1, T = 1)\) in the differential-mode can be achieved when \(\zeta = \left(1, \overline{1}, \overline{1}\right)\). If the performance is calculated for \(\zeta = \left(1, \overline{1}, \overline{1}\right)\) by including the effect of \(\Gamma\) and \(T\) of the two sections [7], the fractional bandwidth is found to be equal to 120%. Using the same approach, it is possible to show that when \(\zeta < \left(1, \overline{1}, \overline{1}\right)\), the fractional bandwidth decreases. The bandwidth increases, but the performance at the center of the passband deteriorates when \(\zeta > \left(1, \overline{1}, \overline{1}\right)\).

For \(\zeta = \left(1, \overline{1}, \overline{1}\right)\), the even \(Z_{ee}\) and odd \(Z_{oe}\) impedances for the two coupled sections are calculated to be \(120.7\) \(\Omega\) and \(20.7\) \(\Omega\), respectively [7]. The required dimensions for the coupled sections \((I^1_2, I^2_2)\) to achieve those impedances can be found from the following relations, which are derived using the conformal mapping technique [7, 10]

\[
Z_{ee} = \frac{\imath \pi \varepsilon_r \ell K(\xi^1_2)}{\sqrt{\varepsilon_r \ell K(\xi^1_2)}}; \ Z_{oe} = \frac{\imath \pi \varepsilon_r \ell K'(\xi^1_2)}{\sqrt{\varepsilon_r \ell K'(\xi^1_2)}} \tag{4}
\]

\[
\ell_1 = \frac{\pi \ell \varepsilon_r \ell K(\xi^1_2) + \imath \pi \varepsilon_r \ell K'(\xi^1_2)}{\sqrt{\varepsilon_r \ell K(\xi^1_2) + \imath \pi \varepsilon_r \ell K'(\xi^1_2)}} \tag{5}
\]

\[
\ell_2 = \frac{\pi \ell \varepsilon_r \ell K(\xi^1_2) + \imath \pi \varepsilon_r \ell K'(\xi^1_2)}{\sqrt{\varepsilon_r \ell K(\xi^1_2) + \imath \pi \varepsilon_r \ell K'(\xi^1_2)}} \tag{6}
\]

\(\lambda_c\) is the effective wavelength at the center frequency (6.85 GHz), \(h\) is the thickness of the substrate, \(K(\xi)\) is the first kind elliptical integral and \(K'(\xi) = K(\sqrt{1 - \xi^2})\).

The dimensions \(\ell_1\) and \(\ell_2\) in Fig. 1(c) are not critical design parameters. However, \(\ell_1\) shouldn’t be too small in order to avoid undesired coupling between the microstrip patch connected to port 1 and that connected to port 1’. The distance between the two sections of the balanced filter, i.e., \(\ell_2\), is chosen just to make enough space for the embedded LPF.

Concerning the embedded LPF, which can be represented by the equivalent RLC circuit in Fig. 1(d), the length of the stubs \((l_d)\) and the slot in the ground plane \((l_s)\) are designed slightly below a quarter of the effective wavelength at the required cutoff frequency. This length ensures that the stubs and slot add effectively pure reactive elements to the microstrip line. For the presented balanced filter, the passband of interest ends at 10.6 GHz. Thus, the cutoff frequency for the embedded LPF is set at 13 GHz. This value ensures that the LPF does not have any negative impact on the passband of the balanced filter, whereas it does have the required impact of reducing the harmonic responses.

### III. RESULTS AND DISCUSSIONS

The designed filter was fabricated using two layers of Rogers TMM4 (\(\varepsilon_{\text{relative}} = 4.4\), \(\tan\delta = 0.0009\)) as the substrate. The design guidelines explained in the previous section were used to calculate the initial dimensions of the devices, whereas the final dimensions were found using the op-
The proposed filter suppresses the common-mode signals across the whole investigated band, i.e., up to 20 GHz. Concerning the UWB range (3.1 to 10.6 GHz), the common-mode signal is attenuated by more than 24 dB according to the simulations and 24 dB in the measurements.

A good agreement can generally be observed between the simulated and measured results depicted in Fig. 3. The slight difference between them can be mainly attributed to the manual alignment and sticking of the different layers used to develop the multilayer structure.

One of the important parameters that defines the quality of the performance of filters is the variation in the group delay across their passband. For a distortionless operation in the pulsed RF systems, the filter is required to maintain flat group delay within its passband. The measured group delay of the developed device depicted in Fig. 3 reveals a flat group delay with only 0.1 ns peak-to-peak variation across the passband.

IV. CONCLUSION

A balanced BPF with UWB performance has been presented. The device utilizes a three-layer broadside-coupled structure. The employed structure behaves as a BPF in the differential-mode and an all-stop filter in the common-mode. The simulated and measured results of the developed device have shown more than 120% fractional bandwidth, sharp and wide upper stopband, and sharp lower stopband. The filter has the capability to suppress the common-mode signals by more than 24 dB.

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Design Method for Ultra-Wideband Bandpass Filter With Wide Stopband Using Parallel-Coupled Microstrip Lines

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Abstract—A method to design microstrip bandpass filters with ultra-wideband (UWB) performance, wide stopband, and practical dimensions is presented. According to the proposed method, three subsections of different lengths and coupling factors are connected to form a stepped-impedance parallel-coupled microstrip structure. A theoretical model is derived and used to find the optimum length and coupling factor for each of those subsections for an UWB passband and suppressed second and third harmonic responses in the stopband. The required performance is realized by generating and proper positioning of three transmission zeros in the upper stopband and three transmission poles in the passband. The theoretical model shows that the total length of the three-subsection coupled structure is one-third of the effective wavelength at the center of the passband. The theoretical model is used to find the required design values for the whole structure. The presented method is validated by building a bandpass filter that has a passband from 3.1 to 10.6 GHz with less than 1-dB insertion loss and a wide upper stopband that extends up to 28 GHz.

Index Terms—Bandpass filter, microstrip filter, ultra-wideband (UWB) filter.

I. INTRODUCTION

Bandpass filters that are based on parallel-coupled microstrip lines are widely used in microwave systems due to their simple structure, low cost, and easy integration with other devices. However, their use in many wideband applications is hindered by two factors. The first one is the presence of the second and third harmonic responses. The undesired second harmonic appears due to the difference in the odd- and even-mode phase velocities of the microstrip coupled structures. That difference becomes larger, and thus, the second-harmonic problem becomes more serious when a substrate with high dielectric constant is used. The third harmonic appears due to the distributed nature of the utilized coupled microstrip lines [1], [2]. The second factor that limits the use of microstrip bandpass filters in wideband systems is the need for a very narrow spacing between the coupled lines for fractional bandwidths of more than approximately 30% [3].

In order to address the aforementioned challenges, a wide range of methods and techniques were proposed [4]–[29]. They include combining parallel-coupled lines with different types and shapes of stubs or multimode resonators [4]–[12], manipulating the shape of the ground structure or the dielectric substrate [13]–[24], and modifying the shape of the coupled lines [25]–[29]. Each of the presented filters in [4]–[29] has its own merits. However, when it comes to the requirements of an easy to manufacture filter that has an ultra-wideband (UWB) passband and an upper stopband that is free from the effect of the second and third harmonics, there are important limitations in [4]–[29]. The upper stopband of the filters presented in [4]–[8], [10], [11], [13]–[15], [17], and [20]–[29] has limited width due to the existence of a strong third harmonic response. The inherent characteristics of the utilized structures in [6] and [7] limits the bandwidth to less than 80% as the passband is bounded by the first and third harmonic responses of the stubs.

The technique used in [9] enables the reduction of the levels of the second and third harmonics, but they are not removed completely. To achieve the required UWB performance in [12] and [16], narrow gaps (0.05 mm) were needed between the coupled lines that have a narrow width of 0.1 mm in [12]. To improve the performance at the upper stopband, a low-pass filter was needed in [18]. The measured results of the filter indicate an insertion loss of 1.8 dB across most of the passband. The performance of a four-section filter in [19] reveals an extended upper stopband, but a limited-width passband.

The conclusion from the aforementioned review is that there is a need for a technique that enables the design of a planar UWB bandpass microstrip filter with less than 1-dB insertion loss across its passband, a wide upper stopband that is free from the presence of the second and third harmonic responses, and a reasonable spacing between the coupled lines. The design presented in this paper endeavours to address that need. To suppress the second and third harmonic responses, the coupled structure is divided into three subsections that have different coupling factors and lengths. A theoretical model is derived for the proposed structure. That model is used to find the required values for the length and coupling factor for each of the coupled subsections so that three transmission poles and three zeros are created in the passband and upper stopband, respectively. It is shown that while the side subsections should be designed to have a loose coupling, the central subsection should have a tight coupling that is almost twice that of the side subsections. The length of the central subsection is also twice that of any of the side sub-
sections. The total length of the coupled structure is one-third of the effective wavelength at the center of the passband.

II. THEORY

Assume that a coupled structure is divided into three subsections, as depicted in Fig. 1. For a symmetrical configuration, the two side subsections are assumed to have the same coupling factor \( c_1 \) and length \( l_1 \). The central subsection has a coupling factor \( c_2 \) and length \( l_2 \). Using the even-odd mode analysis for four-port devices [30], [31], it is possible to show that the reflection \( (S_{11}) \) and transmission \( (S_{21}) \) coefficients of the three-subsection coupled structure are given as

\[
S_{11} = L_1 + B_2
\]

\[
S_{21} = \frac{2}{j\pi} \frac{L_1 + B_2}{1 - \frac{L_1 + B_2}{j\pi}}
\]

\[
e = \frac{B_1 + L_2}{1 - \frac{L_1 + B_2}{j\pi}}
\]

\[
r_s = \frac{1}{\sqrt{1 - \eta_c^2}} \frac{2\pi l_1}{\lambda_c}
\]

\[
s_i = \frac{1}{\sqrt{1 - \eta_c^2}} \frac{2\pi l_2}{\lambda_c}
\]

where \( \lambda_c \) and \( l_i \) are the coupling factor and the length, respectively, of the \( i \)th coupled section, and \( \lambda_c \) is the effective wavelength.

The iterative solution of (1)–(6) for \( S_{21} \) shows that it is possible to generate three transmission zeros at the upper stopband in positions that completely suppress the second and third harmonic responses when the length and coupling factor of the side and central subsections have the following relations:

\[
l_2 = 2l_1 = \frac{\lambda_c}{\eta_c}
\]

\[
c_2 = 2\pi c_1 \left( 1 - \frac{c_1}{\sqrt{2}} \right)
\]

\( \lambda_c \): the effective wavelength at the center frequency of the passband \( f_c \).

The solution (7) reveals that the coupled structure of Fig. 1 should have a total length \( (2l_1 + l_2) \) that is equal to one-third of the effective wavelength at the center of the passband. The solution (8) shows that for a spurious-free stopband, the central subsection having a length, which is twice that of any of the side subsections, should be tightly coupled, whereas the side subsections are loosely coupled.

The performance is calculated using (1)–(8) for wide range of coupling values. In those calculations and the other calculations that follow, it is assumed that the required center frequency for the passband \( f_c \) is 6.85 GHz as the target of the design is a passband from 3.1 to 10.6 GHz. The results are shown in Fig. 2 for different values of the coupling factor \( c_1 \). The corresponding coupling factor \( c_2 \) is calculated using (8).

In all the investigated cases, it is found that the stopband is characterized by the presence of three transmission zeros \( (f_{s1}, f_{s2}, \text{and } f_{s3}) \). The position of those zeros are related to the center of the passband as follows: \( f_{s1} = 2\pi f_c \), \( f_{s2} = 2\pi f_c \), and \( f_{s3} = 2\pi f_c \), as shown in Fig. 2. The frequency \( f_{s2} \) appears at \( 3f_c \) because the total length of the coupled structure, which is equal to one-third of a wavelength at \( f_c \), is equal to one wavelength at \( 3f_c \). Thus, a complete reflection of the signal occurs. The other two transmission zero frequencies \( (f_{s1} \text{ and } f_{s2}) \) are the result of the destructive combination of the direct and coupled/reflected signals at the output port.

Concerning the performance at the passband, it is clear from the results of Fig. 2 that for low values of the coupling factors, there is one transmission pole at the passband. That pole appears exactly at \( f_c \). Increasing the value of the coupling factors results in splitting the central pole into two poles that are symmetrical around \( f_c \). That split appears when \( c_1 \geq 4 \). The two poles move towards the two edges of the passband with the continuous increase in the values of the coupling factors resulting in an increase in the width of the passband.

If two sections similar to that shown in Fig. 1 are connected in series, the effective reflection \( (S_{11}) \) and transmission \( (S_{21}) \) coefficients can be found as follows. Assume that a transmission line that has a characteristic impedance \( (Z_0) \) and length \( (\delta) \) is used to connect the two sections that have a characteristic...
impedance \( Z_e = \chi(\frac{1}{2}) \). The connecting line has the following \( \hat{s} \)-parameters that can be derived from its \( \mathbf{Y} \)-\( \mathbf{T} \)-parameters [3]:

\[
\hat{s}_{11c} = \frac{j \left( \frac{Z_1}{Z_0} - \frac{Z_2}{Z_0} \right) \sin(\hat{s})}{2 \cos(\hat{s}) + j \left( \frac{Z_1}{Z_0} + \frac{Z_2}{Z_0} \right) \sin(\hat{s})} \tag{9}
\]

\[
\hat{s}_{21c} = \frac{2}{2 \cos(\hat{s}) + j \left( \frac{Z_1}{Z_0} + \frac{Z_2}{Z_0} \right) \sin(\hat{s})} \tag{10}
\]

The effective \( \hat{s} \)-parameters of the whole structure (coupled section #1, connecting line, and coupled section #2) can be calculated from the \( \hat{s} \)-parameters of the three parts [32]

\[
\hat{s}_{11} = \hat{s}_{11c} + \frac{\hat{s}_{21c}^2 \hat{s}_{11e}}{1 - \hat{s}_{11c} \hat{s}_{11}} \tag{11}
\]

\[
\hat{s}_{21} = \frac{\hat{s}_{21c}^2 \hat{s}_{21e}}{1 - \hat{s}_{11c} \hat{s}_{11}} \tag{12}
\]

\[
\hat{s}_{11e} = \hat{s}_{11} + \frac{\hat{s}_{21c}^2 \hat{s}_{11e}}{1 - \hat{s}_{11c} \hat{s}_{11}} \tag{13}
\]

\[
\hat{s}_{21e} = \frac{\hat{s}_{21c}^2 \hat{s}_{21e}}{1 - \hat{s}_{11c} \hat{s}_{11}} \tag{14}
\]

\( \hat{s}_{11} \) and \( \hat{s}_{21} \) are given by (1) and (2).

The iterative solution of (9)–(14) shows that the relations between the lengths and coupling factors of the two-section structure should be as in (7) and (8) in order to suppress the second and third harmonic responses in the stopband. This conclusion is valid for any values of \( Z_e \) and \( \hat{s} \). The choice of \( c_1, l_1, c_2, \) and \( l_2 \) according to (7) and (8) results in three transmission zeros at the same positions as those in the single-section device.

As will be thoroughly discussed in the coming section, there are two possible configurations for the connection of the two sections. In one of the configurations, the two sections are connected directly without the need for any connecting line. In this case, the performance can be calculated from (9)–(14) after assuming \( \hat{s} = 0 \). It is found that in order to achieve a passband from 3.1 to 10.6 GHz with less than 1-dB insertion loss and spurious-free stopband, \( c_1 = (1,4) \) and thus, \( c_2 = (0,4) \) from (8). Snapshots of the calculated performance for those values of \( c_1 \) and \( c_2 \) and two other set of values are shown in Fig. 3. The passband performance is similar to that of the single-section device in having the same three transmission zeros. However, the level of attenuation in the stopband of a two-section device is better than that of the single-section device by more than 20 dB, as revealed when comparing Fig. 2 with Fig. 3 for, say, \( c_1 = (1,4) \) and \( c_2 = (0,4) \). Moreover, the rate of cutoff at the lower and upper stopbands is larger in the two-section compared with the one-section structure.

Concerning the performance in the passband, the extensive calculations and the snapshots shown in Fig. 3 indicate that the passband has two transmission poles: \( f_{p1} = 3.7 \) GHz, and \( f_{p2} = 1.7 \) GHz when \( c_1 < c_4 \). For the case \( c_1 < c_1 < c_4 \), a third transmission pole \( f_{p3} \) appears between \( f_{p1} \) and \( f_{p2} \). That pole moves toward \( f_{p2} \) with increasing \( c_1 \). If \( c_1 \) is increased beyond 0.44, a fourth pole starts to appear from the position of \( f_{p2} \) moving toward the lower edge of the passband. That pole moves the low cutoff frequency of the filter to lower values. Moreover, the third pole moves toward the upper edge of the passband, pushing the high cutoff frequency to higher values. Thus, wider passband can be achieved by increasing \( c_1 \), and consequently, \( c_2 \). A significant increase in the width of the passband beyond the UWB requirement of 109% fractional bandwidth comes at a cost; the performance at the center of the passband deteriorates as shown in Fig. 3 for the case \( c_1 = (1,4) \), which achieves 150% fractional bandwidth. The results of Fig. 3 also indicate that by decreasing \( c_1 \), and thus, \( c_2 \), narrowband performance with very high attenuation in the stopband is realized.

When assuming independent values for the coupling factors \( c_1 \) and \( c_2 \), i.e., they are not related according to (8), it is found that the position of \( f_{c2} \) does not depend on the values of \( c_2 \) or \( c_2 \), as revealed in Fig. 4. It is only defined by the value of \( f_c \), and thus, by the length of the coupled structure. On the other hand, the values of the other two transmission zero frequencies \( f_{c1} \) and \( f_{c2} \) depend on the coupling values if they are not taken according to (8). For example, increasing the value of \( c_2 \) to more than the value given by (8) results in an increase in the frequency spacing between the transmission zeros, whereas decreasing that value results in a decrease in the spacing between the zeros. As shown in Fig. 4, both of those two situations cause the unwelcomed appearance of local maxima in the insertion loss at the exact positions of the two side transmission zeros of the case when (8) is applied, i.e., if \( f = 2.7f_c \), and \( f = 3.7f_c \).

In the second configuration of a two-section device, a transmission line of length \( \hat{s} \) and impedance \( Z_e \) is needed to connect the two sections. The performance for this configuration is calculated using (9)–(14) for a wide range of values for \( Z_e \) and \( \hat{s} \). In order to achieve a 3.1–10.6 GHz passband, \( c_1 = (1,4) \), and thus, \( c_2 = (0,4) \) are used in those calculations. Snapshots of the calculations are shown in Figs. 5–7. The results of Figs. 2–7 prove that if \( c_1, l_1, c_2, \) and \( l_2 \) are chosen according to (7) and (8), the three transmission zero frequencies do not depend on the number of sections or parameters of the connecting line.

It is found that if \( Z_e < Z_c \), the best performance is achieved when the connecting line is as short as possible, as revealed in
Fig. 4. Calculated performance of a two-section structure when $c_2 = 0.81$.

Fig. 5. Calculated performance at $c_1 = 0.43$ and $c_2 = 0.81$ for the given length $d$ (per $\lambda_c$) of the connecting line that has $Z_c = 20 \, \Omega$.

Fig. 6. Calculated performance at $c_1 = 0.43$ and $c_2 = 0.81$ for different values of the length $d$ (from 0 to 0.25$\lambda_c$) of the connecting line that has $Z_c = 50 \, \Omega$.

Fig. 7. Calculated performance at $c_1 = 0.43$ and $c_2 = 0.81$ for the given length $d$ (per $\lambda_c$) of the connecting line that has $Z_c = 100 \, \Omega$.

III. DESIGN

From the practical point of view, it is easy to achieve the required loose coupling of 0.43 (equivalent to $-7$ dB) at the two side subsections using conventional parallel-coupled microstrip lines. However, the tight coupling of 0.81 (equivalent to $-1.83$ dB) at the central subsection cannot be realized using the conventional structure. Therefore, two techniques are employed in order to facilitate the achievement of tight coupling in that subsection without the need for a very narrow gap between the coupled lines. Since the tight coupling requires very high even-mode impedance and very low odd-mode impedance, one of the employed techniques aims at decreasing the odd-mode impedance, while the other aims at increasing the even-mode impedance. Both targets are achieved while a reasonable gap between the coupled lines is maintained.

In the first approach, a chip capacitor is connected between the two coupled lines at the middle of the central subsection to increase its odd-mode capacitor, and thus, to decrease its...
odd-mode impedance [33], [34]. That capacitor has no effect on the even-mode circuit. In the second approach, the conductive layer in the ground plane located directly underneath the coupled structure is removed leaving a slotted ground plane at that place. This action results in a reduction in the even-mode capacitor, and thus, an increase in the even-mode impedance.

The proposed method can be implemented using parallel-coupled microstrip lines in two different manners (inline and cascaded configurations), as shown in Fig. 8.

Using the well-known equations that relate the coupling factor with the mode impedances [3], it is possible to show that for the side subsections \((c_1 = 1, 13)\) and the central subsection \((c_2 = 1, 13)\), the even- and odd-mode impedances are equal to 79.2, 31.5, and 154.3, and 16.2 \(\Omega\), respectively. To ease the manufacturing process, the minimum value of the gaps between the coupled lines, i.e., \(s_1\) and \(s_2\) in Fig. 8, is assumed to be 0.2 mm. With \(s_1 = 1.2\) mm and for a certain substrate, the width of the side subsections \((\text{one})\) to achieve the required mode impedances of the side subsections are found using the design equations of the conventional coupled microstrip lines [35], [37].

Concerning the central subsection, the analysis for parallel-coupled lines with slotted ground [35] can be employed after considering the effect of the additional chip capacitor. The even-mode \((Z_{ee})\) and odd-mode \((Z_{oe})\) impedances are equal to

\[
Z_{oe} = \frac{\varepsilon_r K'(l_1)}{\sqrt{\varepsilon_r K''(l_1)}}
\]

\[
Z_{ee} = \frac{\varepsilon_r K'(l_1)}{\sqrt{\varepsilon_r K''(l_1)}}
\]  \hspace{1cm} (15)

\[
K(l) = \frac{\text{first kind elliptical integral and its complementary, respectively, } \varepsilon_r \text{ and } l; \text{ dielectric constant and thickness of the substrate. The design parameter } \nu_2, s_2 \text{ and } \nu_s \text{ are shown in Fig. 8. Since the design (15)–(19) are nonlinear, an iterative procedure was adopted to find the required values of the design parameters. That procedure starts by using the minimum value for } s_2 (0.2 \text{ mm}), \text{ the calculated mode impedances, and a certain value for } \nu_2. \text{ The procedure is repeated until reasonable values for the design parameters are found.}}

Since the effect of the harmonic responses is more serious when the utilized substrate has a high dielectric constant and in order to show the effectiveness of the proposed method, the filter was designed and fabricated using the substrate RT6010 \((\varepsilon_r = 10.2, h = (0, 4.4) \text{ mm})\). Assuming that \(s_1 = s_2 = 1.2 \text{ mm}, \text{ the values of the design parameters are } \nu_1 = 4.4 \text{ mm, } l_1 = 1.1 \text{ cm, } s_1 = 1.2 \text{ mm, } \nu_2 = 4.4 \text{ mm, } l_2 = 2.3 \text{ mm, } \nu_s = 3.7 \text{ mm, } s_2 = 0.3 \text{ mm, and } \nu_2 = 4.1 \text{ mm. For a two-section configuration that needs a connecting line between the two sections, the line has } \nu_{c} = 4.3 \text{ mm, and } \nu = 4.1 \text{ mm.}}

\[
Z_{oe} = \frac{\varepsilon_r K'(l_1)}{\sqrt{\varepsilon_r K''(l_1)}}
\]  \hspace{1cm} (16)

\[
Z_{ee} = \frac{\varepsilon_r K'(l_1)}{\sqrt{\varepsilon_r K''(l_1)}}
\]  \hspace{1cm} (17)

\[
l_1 = \frac{1 + \exp(-\pi(\nu_2 - s_2))}{1 + \exp(-\pi(\nu_2 - s_2 - 2 \nu_2))}
\]  \hspace{1cm} (18)

\[
l_2 = \frac{s_2}{s_2 + 2 \nu_2}
\]  \hspace{1cm} (19)

\(K(l)\) and \(K'(l)\): first kind elliptical integral and its complementary, respectively, \(\varepsilon_r\) and \(l\): dielectric constant and thickness of the substrate.

IV. RESULTS AND DISCUSSIONS

The calculated design values were used in the simulation tool (CST Microwave Studio) to calculate the performance of the filter should those values are used without any optimization. The results shown in Fig. 9 indicate that the passband agrees well with the target. The upper stopband has a sharp cutoff rate at 12 GHz and it is free from the presence of the second and third harmonics. However, the insertion loss at the upper stopband, especially around 18 GHz, needs to be increased. The performance also requires an improvement over part of the passband (around 9 GHz). The level of attenuation at the stopband is below that expected in theory. Those discrepancies between the theoretical and full-wave simulated performance are mainly due to the effect of the parasitic elements and substrate dielectric and conductive losses, which are not considered in the theory. Since the investigated band is extremely wide, it is typical to anticipate that the effect of the nonideal elements used in the simulation is significant.

A close inspection of the performance according to the calculated design values in Fig. 9 shows that the response is to be
shifted slightly upward. Thus, the lengths of the coupled structures should be decreased slightly. The central transmission pole also appears at 7 GHz, whereas it should be at around 8.9 GHz indicating lower achieved values for the coupling factors. Thus, the optimization process should target increasing those factors.

The simulation tool was used to find the optimum dimensions. To ease the manufacturing process, the cost function of the utilized optimization algorithm was designed to drive $s_1$ and $s_2$ to have high values with a minimum acceptable value set at 0.2 mm. Moreover, the added chip capacitor $\varepsilon_x$ was allowed to take only the available standard values. The optimized values are found to be $s_1 = s_2 = 1.44$ mm; $i_1 = i_2 = 0.44$ mm, $l_1 = 1.0$ mm, $l_2 = 1.8$ mm, $i_v = 4.3$ mm, and $\varepsilon_x = 1.0$ pF.

Concerning the inline structure, the connecting line between the two sections was found to have the optimum dimensions $i_x = 0.4$ mm, $d = 1.47$ mm.

In the proposed design, a broadband microwave chip capacitor from Murata Electronics, Kyoto, Japan, was used. As shown in Fig. 8, a tapered microstrip is used to connect the coupled structure to the input and output ports. The use of the tapered line is useful in easing the constraint on the even- and odd-mode impedances of the different subsections. It is found during the optimization that a deviation of about $\pm 10\%$ from the relation $\sqrt{\mu_{\text{even}}/\mu_{\text{odd}}} = \lambda/4$ can be easily compensated using the tapered microstrip lines.

If the calculated values of the design parameters are compared with the optimized values, it is possible to conclude that the presented design method, which is based on the quasi-static assumption, gives reasonable initial values. The major difference occurs in the gap spacing. That difference occurs due to the adopted optimization procedure, which encourages a larger value for the gaps for easy manufacturing.

It is worth mentioning that different optimization techniques were tried. It was found that the best performance is always achieved when the width and the gap of the central subsection $(i_2$ and $s_2)$ are equal to those of the side subsections $(i_1$ and $s_1)$. The other design parameters of the central subsection ($\varepsilon_x$ and $i_v$) are used to get its required mode impedances, which are different from those of the side subsections. This result can be explained by the removal of all the discontinuities from the structure when the side and central subsections have the same width and gap.

Using the optimized design values, the performance of the inline configuration [see Fig. 8(a)] and the cascaded configuration [see Fig. 8(b)] were simulated. The results shown in Fig. 9 reveal that both of the structures achieve the required UWB performance. The use of the optimized parameters results in a significant improvement in the performance, especially at the upper ends of the passband and stopband. Fig. 9 shows that the inline configuration has a better performance concerning the return loss at the passband and the insertion loss at the upper stopband. The return loss of the inline structure is more than 20 dB across the band from 3.5 to 10.6 GHz, whereas the return loss in the cascaded structure is better than 17 dB across the same band. The insertion loss of the inline structure is more than 40 dB, whereas it is more than 31 dB for the cascaded structure, across the stopband from 13.5 to 19.5 GHz.

The improvement in the performance of the inline structure compared with the cascaded structure is achieved by the proper utilization of the connecting transmission line and a weak capacitive end coupling between the first and second sections. The level of that capacitive coupling is controlled using the distance $d$ [see Fig. 8(a)] between the first and second sections of the filter. Concerning the size of those coupled structures, it is easy to show that the inline structure needs a smaller size than that of the cascaded structure under the realistic assumption $d < (l_1 + (\lambda/2))$. For those reasons, the inline structure is chosen for the development of the proposed filter. The overall dimension of the manufactured filter (Fig. 10) is 1.2 cm $\times$ 1.7 cm.

It is worth mentioning that with a smaller size and a better performance of the inline structure compared with the cascaded structure, the parallel-coupled resonators used by Cohn [37] for the development of bandpass filters could be arranged in an inline configuration instead of the originally suggested cascaded configuration. The capacitive end coupling between subsequent sections can then be utilized to improve the performance in a similar manner to the method used in this paper.
Fig. 11. Measured and simulated $S$-parameters for the inline configuration across the investigated band.

Fig. 12. $S$-parameters and group delay for the inline configuration across the lower part of the investigated band.

The simulated and measured performance of the inline configuration is shown in Fig. 11 for the whole investigated band and in Fig. 12 for the passband and small parts of the stopbands. Figs. 11 and 12 reveal a passband from 3.1 to 10.6 GHz with less than 1-dB insertion loss. The return loss is more than 20 dB according to the simulations and 18 dB in the measurements across the band from 3.5 to 10.5 GHz. Within the passband of the filter, there are three resonances that appear at 5, 8, and 10 GHz in the return loss as expected in the theoretical model. The device has a sharp cutoff at the upper stopband. The maximum insertion loss is equal to 48 dB at 15.8 GHz in the simulations and 47 dB at 15.4 GHz in the measured results. The insertion loss is more than 30 dB across the stopband from 12 to 24 GHz. The upper stopband extends up to 28 GHz.

As shown in Fig. 12, the simulated and measured results agree very well across the band up to around 14 GHz. The difference between them is significant after that frequency, as revealed in Fig. 11. That difference can be attributed to two parameters. The first one is the performance of the utilized chip capacitor. Although the capacitor was chosen to be of a microwave broadband type, its performance according to the technical data is guaranteed up to 10 GHz. However, after 10 GHz, it could have parasitic parameters (inductors and resistors) that affect its performance. The second parameter is the Subminiature A (SMA) connectors. The utilized connectors have less than 0.4-dB insertion loss for operating frequencies up to 18 GHz. Beyond that frequency, the insertion loss of each of those connectors increases significantly.

To quantify the level of distortion introduced by any bandpass filter, it is necessary to measure the deviation in the group delay across the passband of the filter. Concerning the developed filter, the measured group delay depicted in Fig. 12 shows a low peak-to-peak deviation of 0.12 ns across the band from 3.1 to 10.6 GHz.

V. CONCLUSION

An UWB bandpass filter based on parallel-coupled microstrip lines has been presented. In order to achieve the required UWB passband and wide spurious-free stopband, three subsections that have different lengths and coupling factors are connected to form a stepped-impedance coupled structure. The resultant structure creates three transmission poles at the passband and three transmission zeros at the upper stopband. A theoretical model is derived and used to find the optimum length and coupling factor for each of those subsections for an UWB passband with suppressed second and third harmonic responses. To achieve those targets, it is shown that the central subsection is required to be tightly coupled; whereas the side subsections are loosely coupled. The derived model shows that the total length of the three-subsection coupled structure is one-third of the effective wavelength at the center of the passband. The presented method is validated by simulations and measurements.

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Lowpass filter utilising broadside-coupled structure for ultrawideband harmonic suppression

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Abstract: A single-substrate microstrip/coplanar waveguide broadside-coupled structure is utilised to build a lowpass filter. The main features of the proposed filter are the extremely wide stopband, sharp cutoff response, compact size and closed-form design procedure. The theory of operation for the proposed filter is presented, its design procedure is derived and its performance is explained. The measured and simulated performance of a manufactured prototype filter agrees well with the theoretical analysis. The filter shows a negligible radiation, flat group delay and a sharp and wide stopband that is larger than 14 times the 3 dB cutoff frequency of the filter. The length of the coupled structure required to build the filter is less than 6% of the effective wavelength calculated at the cutoff frequency.

1 Introduction

With the huge expansion of communication systems, high performance lowpass filters (LPF) have become essential components in those systems. In a typical communication system, mixers and oscillators are usually followed by a low pass filter to remove the harmonic products. A wide stopband, sharp cutoff response and compact size are important factors in the design of the lowpass filter as required by the modern communication systems.

Conventional microstrip lowpass filters such as stepped-impedance filters, open-stub filters, parallel-coupled half-wavelength resonator filters, hairpin-line filters, interdigital and comb-line filters are widely used in many RF applications [1]. However, those filters show generally poor spurious response in their stopband. Some techniques have been introduced to address this serious problem. For the case of open stub LPFs, slotted ground plane [2–9], multi-section structures [10], broadened transmission line elements over a defected ground plane [11–13], short-ended coupled lines [14], resistive loading using thin or thick-film technology [15], and the combined use of open stubs and stepped impedance resonators [16] are utilised to extend the width of the stopband.

In another strategy to widen the stopband of the LPFs, transmission line loaded with slow-wave resonators [17, 18] or tapered stubs [19–22], was investigated. The interdigital structures were also utilised in a multi-section structure [23], or loaded with tapered resonators, to obtain a wide upper stopband [24]. In another approach, the negative permittivity property of the complementary split ring resonators was used [25–29]. The combination of interdigital structures and slotted ground plane was also proposed [30].

The lowpass filter with high–low impedance, or stepped impedance lines, has limited applications in its basic structure because of the narrow stopband and a poor cutoff response. To improve the performance of the stepped impedance lines, different techniques were proposed. It is shown that the use of stepped-impedance hairpin resonators [31, 32], defected ground [33–36], tapered shunt stubs [37, 38], finite ground plane [39], folded resonators on liquid–crystal–polymer substrate [39], interdigital structures [40–42] or coupled-line hairpin unit [43, 44] is useful in extending the stopband of the LPFs.

Reviewing the performance of the LPFs designed using the techniques explained in [2–45] shows that the stopband is extended significantly compared with the conventional structures of microstrip LPFs. The ratio of the harmonic frequency to the cutoff frequency of the filter is extended to be in the range between four and nine [7, 8, 17, 20, 21, 24, 30, 32–34, 36, 38, 41, 42, 44]. However, some of the utilised techniques result in a large size [3, 5, 14, 19, 21, 29, 31] or high insertion loss at the passband [5, 6, 15, 37].

The performance of some of the proposed LPFs shows high radiation losses at parts of the stopband [12, 21, 23, 28, 31, 37]. This phenomenon casts some doubts on whether the designed filters do their job normally in reflecting the signals that have frequencies within the stopband, or behave as a radiator at parts of the assumed stopband. Some of the presented filters rely on very narrow coupled slots [4, 17, 24, 37, 38, 43] or very narrow transmission lines [14, 34]. In other filters [2, 5, 6, 9, 11–13, 15, 16, 26–28, 31, 39, 40, 43, 45], the performance is shown across a limited-width stopband, and thus, it is not possible to verify the full extension of the stopband.

Recently, it has been shown that broadside-coupled microstrip/coplanar waveguide (CPW) structures can be utilised to build microwave devices with high-quality performance [46, 47]. The performance of those devices can be controlled over a wideband using the coupling factor [46]. One of the important characteristics of those
broadside-coupled structures is the ability to achieve tight coupling that is almost constant across a wideband. The presented method in this paper exploits the strong electromagnetic broadside coupling between a microstrip patch at the top layer of a substrate and a CPW at the bottom layer of the substrate to build an LPF with an extremely wide stopband. A complete design method is presented and verified via simulations and measurements. It is shown that the proposed filter has a compact size, sharp cutoff and an extremely wide stopband.

2 Theory

A general coupled structure is assumed to be connected in the manner shown in Fig. 1a. One of the coupled elements is grounded at the centre and it is open-ended at both sides, whereas the other coupled structure is connected to the input/output ports. The structure of Fig. 1a can be analysed by considering it as a two-section four-port coupler as revealed in Fig. 1b. It is assumed that the coupling factor is equal to $C$ between the coupled patches and that the input and output ports are perfectly matched. Depending on the analysis of four-port couplers [48], it is possible to show that the reflection ($S_{11}$) and transmission ($S_{21}$) coefficients are equal to

$$S_{11} = B^2 \frac{1 - A^2 + B^2}{1 + A^2 + B^2} \tag{1}$$

$$S_{21} = A^2 \frac{1 + A^2 - B^2}{1 + A^2 + B^2} \tag{2}$$

$$A = \frac{\sqrt{1 - C^2}}{\sqrt{1 - C^2} \cos(\beta_{ef} l/2) + j \sin(\beta_{ef} l/2)} \tag{3}$$

$$B = \frac{jC \sin(\beta_{ef} l/2)}{\sqrt{1 - C^2} \cos(\beta_{ef} l/2) + j \sin(\beta_{ef} l/2)} \tag{4}$$

where $l$ is the physical length of the coupled structure and $\beta_{ef}$ is the effective phase constant in the medium of the coupled structure.

Using (1)–(4), it is possible to show that $A = 0$, $B = 1$, $S_{21} = 0$, $S_{11} = 1$, and thus, the structure of Fig. 1 behaves as an all-stop filter when $C = 1$, whereas it behaves as an all-pass filter when $C = 0$. This means, theoretically at least, that it is possible to use the coupling factor to achieve a stopband of any width between zero, when $C = 0$ and infinity when $C = 1$.

If the performance is calculated using the above derived equations for different values of the coupling factor, the result is as shown in Fig. 2. It is clear from Fig. 2 that in order to design an LPF with an extremely wide stopband, a tight coupling is required. This conclusion is verified by plotting the relation between the width of the stopband, as defined in Fig. 2 and the coupling factor using the data from (1) to (4) and the result is shown in Fig. 3.

In order to simplify the design procedure for the LPF, curve fitting is utilised to show that the relation between the width of the stopband normalised by the value of the cutoff frequency ($f_c$), that is $W_{sb} = \text{width of the stopband}/f_c$, and the coupling factor is

$$W_{sb} = 0.05e^{4.8C} + 5 \times 10^{-10}e^{24.3C} \tag{5}$$

There is an excellent agreement between the values of the...
width of the stopband calculated using (5) and those obtained using the derived method (1)–(4) as shown in Fig. 3. The above equation is valid for any substrate with any thickness.

The results shown in Figs. 2 and 3 reveal that it is possible to design an LPF with an extremely wide stopband by choosing the design parameters such that the coupling factor for the structure shown in Fig. 1 is close to 1. If a broadside-coupled structure is used, the tight coupling can be easily achieved as proven by different ultrawideband devices that have tight coupling [46, 48]. The utilised structure in this paper is the microstrip/coplanar waveguide-coupled structure shown in Fig. 4. It can be designed using a two-sided printed circuit board. The top layer includes a microstrip patch that is connected with the input/output ports. The bottom layer contains the ground plane. There is a coplanar waveguide at the centre of the ground plane. The tight coupling of this structure is because of a strong electromagnetic coupling between the microstrip patch located at the top layer and the CPW at the bottom layer. To make the structure fully equivalent to the theoretical model of Fig. 1, the CPW at the bottom layer is grounded at the centre using a microstrip line as depicted in Fig. 4b.

It is worth mentioning that the structure of Fig. 4 is adopted in the proposed filter because it is easy to manufacture when compared, for example, with the multilayer microstrip-slot-microstrip configurations [48]. Moreover, the short circuit needed between the bottom-layer coupled structure and the ground plane can be easily implemented in the utilised structure of Fig. 4 without the need for any bias.

Concerning the length of the coupled structure, it can be shown using (1)–(4) and from Fig. 2 that the frequency of transmission zero occurs when the length of the coupled structure is equal to half of the effective wavelength calculated at that frequency. The transmission zero frequency according to Fig. 2 is

$$f_z = f_c(W_{sb}/2 + 1) \quad (6)$$

Thus, it is possible to show that if the LPF is designed to achieve a normalised stopband width $W_{sb}$, the length of the coupled structure $l$ as a function of the effective wavelength $\lambda_e$ calculated at $f_c$ and $W_{sb}$ is

$$l = \frac{\lambda_e}{2 + W_{sb}} \quad (7)$$

To give an indication of the required size of the proposed LPF, assume that it is required to design an LPF with extremely wide stopband such that $W_{sb} = 15$. The length of the coupled structure in this case is equal to $\lambda_e/17$ revealing a compact size for the proposed filter.

It is clear from (7) that there is no compromise between the requirement for an extremely wide stopband and the small dimensions. Both targets can be realised simultaneously if a suitable configuration that can achieve a tight coupling is utilised. However, an utmost care should be taken when using an extremely small structure ($l \ll \lambda_e/20$) to achieve, say, an infinite wide stopband as the structure in this case does not behave as an ideal backward coupled structure, and thus, the analysis (1)–(7) is not valid for this case.

3 Design

In order to establish the validity of the proposed method and to clarify the design procedure, an LPF with the following characteristics is designed, fabricated and tested. The LPF has a cutoff frequency $= 2$ GHz with the capability to remove the harmonic frequencies up to at least $30$ GHz. Thus, the width of the stopband is $28$ GHz and $W_{sb} = 14$.

The first step in the design is to find the required coupling factor to achieve the abovementioned characteristics. According to the design (5), the required coupling factor is $C = 0.97$. The second step in the design is to determine the dimensions of the coupled region that give this coupling factor. As the structure shown in Fig. 4 is an asymmetrical coupled structure (microstrip/CPW), the conformal mapping techniques for the C- and $\pi$-modes of the structures can be utilised to find the relation between the C- and $\pi$-mode impedances and the physical dimensions of the structure. The other required relation between the C- and $\pi$-mode impedances and the coupling factor is given in [46].

Assuming a quasi-transverse electromagnetic propagation, the C- and $\pi$-mode impedances for the microstrip patch at the top layer ($Z_{mc}$ and $Z_{mn}$) and the CPW at the bottom layer ($Z_c$...
and $Z_{cc}$ can be found using conformal mapping [46]

$$Z_{mc} = \frac{60\pi Kr(k_1)}{\sqrt{\varepsilon_r K'(k_1)}}$$ (8)

$$Z_{cc} = \frac{60\pi K'(k_2)K'(k_3)\sqrt{\varepsilon_r}}{(\varepsilon_r - 1)K(k_2)K(k_3) + K'(k_2)K(k_3)}$$ (9)

$$Z_{m\pi} = \frac{120\pi hK(k_1)}{\sqrt{\varepsilon_r}[2hK'(k_1) + (w_m + w_c)K(k_1)]}$$ (10)

$$Z_{c\pi} = \frac{60\pi \sqrt{\varepsilon_r}}{(\varepsilon_r - 1)K(k_2)K(k_3) + 2K(k_1)K'(k_3) + \varepsilon_r(w_m + w_c)/(2h)}$$ (11)

$$k_1 = \frac{\sinh(\pi w_c/(4h))}{\sqrt{\sinh^2(\pi w_c/(4h)) + \cosh^2(\pi w_m/(4h))}}$$ (12)

$$k_2 = \frac{\sinh(\pi w_m/(4h))}{\sinh(\pi w_c/(4h))}$$ (13)

$$k_3 = w_c/w_s$$ (14)

In (8)–(14), $K$ and $K'$ are the first kind elliptical integral and its complementary, respectively, $w_m$ (width of the microstrip coupled patch at the top layer), $w_c$ (width of the central line of the CPW at the bottom layer) and $w_s$ (total width of the CPW at the bottom layer) are the dimensions shown in Fig. 4, $\varepsilon_r$ is the dielectric constant of the substrate, and $h$ is the substrate thickness.

Using the presented design procedure (7)–(14) and the relation between the C- and $\pi$-mode impedances and the coupling factor [46], the following values are found for the dimensions of the coupled structure assuming the use of the substrate Rogers RO4003C with $\varepsilon_r = 3.55$ and $h = 0.508$ mm: $l = 6.4$ mm, $w_m = 5.8$ mm, $w_c = 6.1$ mm and $w_s = 15.8$ mm.

The whole structure is optimised using CST Microwave Studio and the final dimensions are found to be $l = 7.2$ mm $w_m = 5.45$ mm, $w_c = 5.81$ mm, $w_s = 13.46$ mm, width of the microstrip feeders $w_f = 1.2$ mm, width of the microstrip grounding line at the bottom layer $w_l = 0.56$ mm, and width of the slot in the CPW $s = 0.8$ mm. By comparing the calculated and optimised values of the design parameters, it can be concluded that the difference between them is relatively small reflecting the reliability of the derived design procedure.

The theoretical analysis (1)–(14) assumes that the utilised coupler has an infinite directivity, whereas the practical structure of Fig. 4 is expected reasonably to have a limited directivity. Thus, the following final step of the design is adopted in order to bridge the gap between the directivity of the theoretical model and that of the practical structure. The microstrip coupled patch at the top layer is extended slightly in length (by much less than $\lambda_\nu/20$) compared with the length of the coupled structure at the bottom layer. This action adds an inductive element in series with the coupled structure [49]. As an indication of the extended length, it is possible to make it less than. It was proven that series inductors can be utilised for directivity enhancement of coupled structures [50]. Concerning the filter presented in this paper, the improvement in the directivity of the utilised coupler is transformed into an improvement in the overall performance, that is, the insertion and return losses across the whole investigated band. The optimum length of the added microstrip patch to each side of the coupled microstrip patch at the top layer was found to be 2.7 mm.

4 Results and discussions

The designed low pass filter was manufactured and tested. The developed device is shown in Fig. 5. The simulated and measured scattering parameters of the filter at the frequency band up to 30 GHz are depicted in Fig. 6. In order to show details of the performance at the passband and the cutoff response, the scattering parameters for a limited lower part of the investigated band are shown in Fig. 7, which also shows the group delay of the filter is shown in Fig. 7.

It is clear from the results of Figs. 6 and 7 that the proposed device has a cutoff frequency of 2 GHz as required in the design. The results in Fig. 7 reveal a sharp cutoff across the frequency band that extends from the cutoff frequency (2 GHz) to 2.5 GHz. The filter has a wide stopband that extends up to 29 GHz as depicted in Fig. 6. Thus, it has the ability to remove any harmonic response that may appear across that band. The insertion loss of the filter across the

Fig. 5 Views of the manufactured LPF
a Top
b Bottom

Fig. 6 Scattering parameters of the filter at the investigated frequency band
passband is less than 0.5 dB. At the stopband, the filter attenuates the undesired harmonic responses by more than 20 dB across most of the stopband. The first harmonic response appears at 29.8 GHz according to the simulations and at 29 according to the measurement. Those results agree well with the design requirements where the harmonic response is planned to appear at 30 GHz.

It is to be noted that according to the results shown in Fig. 6, the simulated and measured results agree well up to about 15 GHz. At the frequency band above 20 GHz, there is a significant difference between the simulated and measured results. The main factor behind that difference is the S-miniature A (SMA) connectors. According to the technical specifications of the utilised SMA connectors, they should operate well with a maximum of 0.4 dB insertion loss for each connector up to 18 GHz. However, the loss increases significantly above that frequency. A separate test on those connector shows that it introduces an insertion loss of about 1 dB across the band 20–30 GHz. Thus, the measured return loss across the band from 20 to 30 GHz is more than the simulated return loss by about 2 dB on average.

It is worth mentioning here that the utilised structure does not behave as a radiant element across most of the frequency range of interest because of the small electrical size of the utilised coupler. By removing the effect of the SMA connector’s losses from the measured results, it is found that the radiation losses are less than 0.5 dB at frequencies up to 15 GHz. However, significant radiation occurs at the upper end of the investigated band, specifically around the harmonic response.

The low level of radiation from the proposed filter and its compact size mean that the device can be packaged, if required, inside a small enclosure without any significant impact on its performance. It has been shown via parametric simulations that a minimum distance of 10 mm between the slotted ground of the proposed filter and the lower metallic structure of the enclosure is enough to guarantee the required performance of the filter as revealed in the simulation results with enclosure depicted in Fig. 6.

The other important parameter that defines the quality of performance of the filters is the group delay. The filter should have a flat group delay across its passband so that it adds a minimum or no distortion to the signals especially in the pulsed communication system. The proposed filter was tested for this parameter. The measured results shown in Fig. 7 reveal a flat group delay across the passband with the peak-to-peak variation in the group delay of less than 0.2 ns.

If the proposed filter is compared with the other lowpass filters that are designed to achieve the same target of having a wide stopband [2–45] it is possible to claim that the presented filter has a more compact structure, simpler topology and wider stopband. The presented filter is designed using an easy-to-manufacture structure following a closed-form procedure and thus, it saves the time and effort needed in the trial-and-error approach followed by many of the previously proposed structures.

5 Conclusion

A broadside-coupled microstrip/coplanar waveguide has been utilised to design a low pass filter with extremely wide stopband. A theoretical model for the proposed structure has been used to derive a closed-form design procedure for the filter. The simulated and measured results of a prototype filter, designed using the presented method, agree well with each other and with the theoretical model. The manufactured filter has an extremely wide stopband, compact size, negligible radiation and flat group delay.

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Compact tunable low-pass filter using variable mode impedance of coupled structure

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Abstract: A compact and tunable low-pass filter utilising a coupled structure is presented. A closed-form design method is used to estimate the required design parameters so that the filter can have a tunable cutoff frequency across wide range of values. The filter also has an extremely wide stopband, sharp cutoff response and compact size. The theory of operation for the proposed filter is explained, and its design procedure and equivalent circuit are derived. The tuning capability of the filter is enabled by using two varactor diodes connected between the coupled lines to change their $\pi$-mode impedance. The simulated and measured results of a prototype with dimensions 15 mm $\times$ 20 mm designed to have a cutoff frequency tuning range from 1.5 to 2.5 GHz and a stopband that extends to 25 GHz prove the validity of the presented design method.

1 Introduction

Low-pass filters (LPFs) that have wide stopbands are vital in modern wireless systems. In a typical system, mixers and oscillators are usually followed by a LPF to remove the harmonic products. A wide stopband, sharp cutoff response and compact size are important factors in the design of high-performance LPFs.

Conventional microstrip LPFs using stepped-impedance, open-stub, parallel-coupled half-wavelength resonator, hairpin-line and comb-line structures show generally poor spurious response in their stopband [1]. Different techniques, such as inter-digital structures, slotted ground, tapered resonators, metamaterials or coupled hairpin-lines, were introduced to address that serious problem [2–8]. Those techniques are effective in extending the width of the stopband to up to ten times the cutoff frequency of the filters.

In order to make the cutoff frequency of the LPF tunable, several methods were proposed [9–12]. The mechanical approach results in a bulky and narrow stopband filter [9], whereas the electrical approach using several pairs of micro electro-mechanical switch (MEMS) switching elements in the slotted ground shows a performance characterised by a narrow stopband [10]. The use of thin-film technology to develop tunable FPFs results in compact structures [11, 12]. However, the main drawbacks in using that technology are the high-insertion loss across the passband and the high cost. For example, the design proposed in [11] has more than 10 dB insertion loss at the passband, whereas the new generation of commercial tunable LPFs has more than 3 dB insertion loss [12].

In this paper, a tunable compact broadside-coupled microstrip/coplanar waveguide (CPW) is utilised to build an LPF with an extremely wide stopband. The filter is based on the broadside-coupled microstrip to CPW structure [8]. Varactor diodes connected between the coupled lines are used to change the $\pi$-mode impedance, and thus, to tune the passband of the device. A complete design method is presented to enable the proper choice of the values of the design parameters, so that the performance indicators of the filter (return loss and insertion loss) stay within the acceptable limits across the tuning range of the filter. The proposed method is verified via simulations and measurements.

2 Proposed device

The proposed LPF is depicted in Fig. 1. It can be explained using two methods. The first one assumes the filter an asymmetrical broadside-coupled structure [8, 13]. This method is used to find the physical dimensions of the device for a certain cutoff frequency. In that method, the top layer is considered to have a microstrip line that is coupled to a CPW at the bottom layer. The central section of the CPW is grounded at the centre using a microstrip line. The second method assumes the structure as a wide microstrip line over a defected ground structure. From that method, the equivalent circuit can be easily derived and used to explain the performance.

The equivalent coupled structure for the proposed device of Fig. 1 is shown in Fig. 2. The coupled line 1 represents the microstrip at the top layer, whereas coupled line 2 represents the CPW at the bottom layer. Two varactor diodes are connected between the two coupled lines symmetrically around the central point. The biasing circuit needed for those diodes is shown in Fig. 1. Two DC-blocking capacitors and a radio frequency choke (RFC) are used to isolate the biasing circuit from the microwave signal.

The structure of Fig. 2 is analysed by considering it as a two-section four-port asymmetrical coupler. Using the signal flow diagrams of four-port devices as depicted in
Fig. 1  Utilised structure of
a Top layer
b Bottom layer

Fig. 2, it is possible to show that for arbitrary coupled structure, the outgoing signals from the four terminals of each of the two sections are given as [13]

\[ b = (I - S\Gamma)^{-1}Sa \]  

where \( b \): 4 \times 1 vector representing the signal out of each terminal. The elements in this vector are given the designations (\( b'_i \)), \( i \) refers to the section number, \( j \) refers to the terminal number, \( I \) is the 4 \times 4 identity matrix and \( a \) is the 4 \times 1 vector of input signals from outside sources.

The elements of \( a \), that is, \( a'_j \), for the structure of Fig. 1b are

For section #1:\n\[ a'_1 = 1; \quad a'_2 = b'_1; \quad a'_3 = a'_4 = 0 \]  

For section #2:\n\[ a'_1 = b'_2; \quad a'_2 = a'_3 = a'_4 = 0 \]

\( \Gamma \): 4 \times 4 matrix representing the reflection coefficients at the four terminals. All the elements of the matrix are equal to zero except the diagonal elements (\( \Gamma'_{jj} \)) with the following values

For section #1: \( \Gamma'_{11} = \Gamma'_{22} = 0; \quad \Gamma'_{13} = 1; \quad \Gamma'_{44} = -1 \)  

For section #2: \( \Gamma'_{11} = \Gamma'_{22} = 0; \quad \Gamma'_{13} = -1; \quad \Gamma'_{44} = 1 \)

\( S \): 4 \times 4 scattering matrix with elements that are calculated using the \( \pi-C \) modes approach as explained in [14]. The main parameters required in that approach are the \( C \)- and \( \pi \)-mode impedances for the microstrip patch at the top layer (\( Z'_{o\alpha} \) and \( Z'_{o\beta} \)) and the CPW at the bottom layer (\( Z'_{oc} \) and \( Z'_{o\pi} \)). The relation between them and the physical dimensions of the utilised structure can be derived using the conformal mapping technique [15] after including the effect of the varactors’ capacitance (\( C_v \))

\[
Z'_{oc} = \frac{60 \pi K(k_1) \sqrt{\varepsilon_r}}{\sqrt{\varepsilon_r K'(k_1)}}; \quad Z'_{o\pi} = \frac{60 \pi K'(k_2) K'(k_3) \sqrt{\varepsilon_r}}{(e_r - 1)K(k_2)K(k_3) + K'(k_2)K(k_3)}
\]

\[
Z'_{o\alpha} = \frac{120 \pi h K(k_1)}{\sqrt{\varepsilon_r[2hK'(k_1) + (w_m + w_c)K(k_1) + 4hC_vK(k_1)/(e_r e_s l)]}}
\]

\[
Z'_{o\beta} = \frac{60 \pi \varepsilon_r}{K(k_2)} + \frac{2K(k_1) + e_r (w_m + w_c) + 2C_v}{2h} e_s f
\]

\[
k_1 = \sqrt{\sinh^2(\pi w_c/(4h)) + \cosh^2(\pi w_m/(4h))}
\]

\[
k_2 = \frac{\sinh(\pi w_c/(4h))}{\sinh(\pi w_m/(4h))}; \quad k_3 = w_c/w_s
\]

\( K \) and \( K' \) are the first kind elliptical integral and its complementary, \( w_m \), \( w_c \) and \( w_s \) are the physical dimensions shown in Fig. 3, \( e_r \) and \( h \) are the dielectric constant and thickness of the substrate, and \( l \) is a total physical length of the coupled structure.

The important parameters that define the performance of the filter as shown in Fig. 1 are the reflection coefficient at the input and output ports (\( S_{AA} \) and \( S_{BB} \)), and the transmission coefficient from the input to the output port (\( S_{BA} \)). Those parameters can be calculated after solving (1)–(5) as follows

\[
S_{AA} = S_{BB} = b'_1/a'_1; \quad S_{BA} = b'_2/a'_1
\]

The derived model (1)–(11) is included in an iterative Matlab code aimed at finding the optimum performance for different dimensions of the coupled structure. For a tunable range of cutoff frequency extending, for example, from 1.5 to 2.5 GHz, it is found that the required design parameters assuming the use of the substrate Rogers RO4003C with \( e_r = 3.55 \) and \( h = 0.508 \) mm are: \( w_m = 3.8 \) mm, \( w_c = 2.5 \) mm, \( w_s = 11.55 \) mm, \( l = 7.5 \) mm, and \( C_v \) ranges from 0.4 to 1.4 pF. The above results indicate a compact device as the length of the coupled structure is less than one-tenth of the effective cutoff wavelength.
3 Results and discussions

The performance of the designed device with the calculated dimensions is simulated as depicted in Fig. 3. The filter has a tunable cutoff frequency extending from 1.6 to 2.4 GHz with an ultrawideband stopband extending to more than 25 GHz. The other main feature of the reported performance is a sharp cutoff owing to a transmission zero at around 3 GHz. However, the attenuation at the stopband is around 15 dB, and thus, it requires an improvement.

To explain the resultant performance of the filter and to explore methods to improve the attenuation at the stopband, the equivalent circuit of the filter is derived. This target can be achieved either from the simulated S-parameters or by looking at the proposed structure from another viewpoint that complements the previously explained perspective of the device as a coupled-structure. The structure of Fig. 1 can be assumed a wide microstrip line over a dual face-to-face C-shaped defected ground structure. Following [16], the interaction between the microstrip line and each of the C-shaped slots in the ground is equivalent to a parallel LC-resonator that represents a one-pole Butterworth filter. The wide microstrip line connecting the two sections can be represented as a shunt capacitance. Thus, the simple equivalent circuit representation of the filter including the capacitances of the two varactor diodes is shown in the dotted box of Fig. 4. This circuit can be simplified into a three-order filter. As the simulated passband response of the filter depicted in Fig. 3 is similar to that of Butterworth filters, the values of the different elements of the circuit can be derived from a three-order Butterworth prototype using suitable frequency and impedance scaling.

One of the important points that become clear from the equivalent circuit is that the parallel LC resonator is responsible for the transmission zero in the performance depicted in Fig. 3 at 3 GHz. Thus, it is possible to say that $1/(2\pi\sqrt{L_1C_1}) = 3$ GHz. To change the position of that zero, $L_1$ and/or $C_1$ have to be changed, and thus, the dimensions of the C-shaped slot have to be changed. From the coupled-method perspective, this is equivalent to changing the dimensions of the CPW at the bottom layer.

In order to increase the attenuation at the stopband, the order of the filter is to be increased. One of the possible simple solutions is to add two shunt capacitors ($C_3$) in the manner shown in Fig. 4 resulting in a five-order filter. Those two capacitors can be realised by extending the wide microstrip line at the top layer. For a Butterworth response with a 3 dB cutoff frequency at 2 GHz and better than 25 dB attenuation at the stopband, it is possible to obtain a rough estimation of the required shunt capacitance $C_3$ as 0.9 pF from the equivalent five-port Butterworth prototype. The required extended length ($l_s$) of the wide microstrip line can then be calculated from the value of the capacitance $C_3$. The relation between them is [17]

$$l_s = \left(\frac{\lambda_{cc}}{2\pi}\tan^{-1}\left(2\pi C_3 Z_{om}\right)\right)$$

$Z_{om}$ is the characteristic impedance of the extended wide microstrip line, and $\lambda_{cc}$ is the effective wavelength of the extended microstrip line at the central cutoff frequency (2 GHz). Using the calculated design values from the coupled-structure method, it is possible to find that $Z_{om} = 18$ $\Omega$, $\lambda_{cc} = 8.5$ cm and thus $l_s = 2.7$ mm. The values of the other elements of the equivalent circuit of Fig. 4 are $L_1 = 3.58$ nH, $C_1 = 0.79$ pF and $C_2 = 3.2$ pF. Using the simulator CST Microwave Studio, the performance of the modified structure with the calculated design values depicted in Fig. 3 indicates a significant increase in the stopband attenuation from 15 to 25 dB. However, the tunable cutoff frequency extends from 1.7 to 2.7 GHz.

The simulator CST is used to optimise the dimensions for a tunable cutoff frequency from exactly 1.5 to 2.5 GHz with the

![Fig. 3](image-url)  
*Fig. 3 Performance of the initial and modified structures using the calculated design values*

![Fig. 4](image-url)  
*Fig. 4 Equivalent circuit of the proposed filter*
The smallest possible structure. The optimised values (mm) of the design parameters are: \( w_m = 3.5 \), \( w_c = 2.3 \), \( w_s = 12.2 \), \( l = 7.9 \), \( l_s = 1.8 \), \( d = 0.5 \), \( s = 0.7 \) and \( C_v = 0.3 \) to 1.5 pF.

A prototype of the optimised filter is manufactured as depicted in Fig. 5. It has a compact size of overall dimensions 15 mm × 20 mm. The performance of the filter is verified via simulations and measurements after manufacturing a prototype (Fig. 5). Two GaAs hyperabrupt junction varactor diodes with a maximum biasing voltage of 12 V and total capacitance that covers the required \( C_v \) range are used. To effectively isolate the biasing circuits from the microwave signal path, two 100 nF DC-blocking capacitors and RFC of 10 \( \mu \)H are used. It is worth mentioning that high value for the coupling capacitors is used to keep the lower end of the passband close to DC. In the current design, the lower end of the passband is at 0.02 MHz.

The simulated and measured performance of the filter at the frequency band up to 25 GHz and for different biasing states of the diodes is depicted in Fig. 6. It is clear that the proposed device has a tuning range for the 3 dB cutoff frequency from 1.5 to 2.5 GHz in the simulation and 1.25 to 2.3 GHz in the measurement. The slight difference in the tunable range of cutoff frequency between the measured and simulated results is because of the diodes’ effective series resistance and inductance, and packaging capacitance. The results also show a sharp rate of cutoff at all the diodes’ biasing states with a transmission zero at about 3 GHz. The filter has a wide stopband that extends to more than 25 GHz.

The results in Fig. 6 show a significant difference between the simulated and measured return loss after the frequency 12 GHz. This can be attributed to the utilised sub-miniature A connectors at the input/output ports and the non-ideal performance of the utilised substrate at such high frequencies.

**4 Conclusions**

A compact coupled structure with variable mode impedance has been used to design an LPF that has an extremely wide stopband and tunable passband. Two varactor diodes connected between the coupled lines are used to change the cutoff frequency. The simulated and measured results of a prototype designed to have a 3 dB tunable cutoff frequency
from 1.5 to 2.5 GHz and a stopband that extends to 25 GHz prove the validity of the presented method.

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Ultra-Wideband Phase Shifters

Amin M. Abbosh

Abstract—A method with clear guidelines is presented to
design compact planar phase shifters with ultra-wideband (UWB)
characteristics. The proposed method exploits broadside coupling
between top and bottom elliptical microstrip patches via an
elliptical slot located in the mid layer, which forms the ground
plane. A theoretical model is used to analyze performance of the
proposed devices. The model shows that it is possible to design
high-performance UWB phase shifters for the 25°–48° range
using the proposed structure. The method is used to design 30°
and 45° phase shifters that have compact size, i.e., 2.5 cm × 2 cm.
The simulated and measured results show that the designed phase
shifters achieve better than ±3° differential phase stability, less
than 1-dB insertion loss, and better than 10-dB return loss across
the UWB, i.e., 3.1–10.6 GHz.

Index Terms—Aperture coupling, phase shifter, ultra-wideband
(UWB).

I. INTRODUCTION

PHASE shifters are common microwave devices, which are
widely used in electronic beam-scanning phased arrays,
microwave instrumentation and measurement systems, modu-
lators, and many other industrial applications. In these and
many other applications, and in order to get wideband perfor-
ance, phase shifters are usually realized in planar (stripline or
microstrip) technology due to its nondispersive and broadband
propagation properties.

In order to achieve the broadband operation of the phase
shifters, the approach of coupled transmission lines is usually
employed. One of the earliest designs that used the coupled lines
method to construct broadband phase shifters is the Schiffman
differential phase shifter [1]. It consists of two transmission
lines: the reference transmission line and the folded edge-cou-
pled section. By the proper selection of the length of these
lines and the degree of coupling, the phase difference between
them can be made constant at 90° over a broadband. However,
Schiffman’s original study was based on stripline transmission
structures, where the odd and even modes propagating along
the coupled lines have equal phase velocities. When this type
of circuit is designed in a microstrip form, the unequal odd-
and even-mode velocities results in poor performance [2].
Moreover, the measured results of Schiffman’s phase shifter
indicate a high phase ripple (±10°) [1].

In order to obtain a broader bandwidth with an acceptable
phase ripple using the edge-coupled method, some authors
proposed the use of cascaded multiple coupled sections [3]–[7].
Shelton and Mosko [4] described an approximate synthesis
technique for fixed phase shifters consisting of multiple
parallel-coupled quarter-wave sections. The main drawback
of this procedure, which is general to edge-coupled shifters,
is that in order to achieve a broadband, an extremely tight
coupling is required, which may not be realizable in a given
practical configuration. Shelton and Mosko [4] proposed the
use of tandem coupling to minimize the effect of this problem.
However, tandem configuration requires wire crossovers,
which is inconvenient from the manufacturing point of view.
The designed configuration in [4] suffers from another serious
problem, which is the large size required for the multiple
coupled sections.

In order to decrease the size of the multisection coupling
structure required in the design presented in [4], an optimization
technique was used to calculate parameters of the edge-coupled
broadband phase shifter [5]. The structure considered in [5] con-
ists of a cascade of coupled line pairs of varying length and cou-
pling coefficients and each connected together at one end. The
main drawback of the adopted technique is that it still requires
a large number of coupled line pairs to achieve the required phase
performance.

Some other modifications were proposed to improve perfor-
mance of the edge-coupled structures using new forms of mul-
tisection coupling lines [6] and a double or parallel Schiffman
phase shifter [7]. However, the design presented in [6] requires
a very narrow slot (tens of micrometers), which makes the fab-
rication process very difficult. Moreover, the measured phase
performance presented in [6] shows a phase shifter that covers
only the (4–8 GHz) with a phase error of ±6.5°. The cir-
cuit presented in [7] (referred to as a double parallel Schiffman)
consists of two parallel-connected coupled sections designed to
yield a 90° phase difference. The lengths and coupling values
are adjusted to obtain a desired phase ripple. The measured re-
results of the design in [7] indicate a narrowband performance
with high phase instability at the lower and higher ends of the
frequency band.

A compact version of the Schiffman phase shifter was intro-
duced in [8]. Although the proposed design uses a smaller area
and it is a cost-effective one compared with the original design,
it has a narrow bandwidth.

Tresselt explained a different design procedure using a
continuously tapered coupled section [9]. He noticed that
the spread in coupling values between adjacent sections of
the cascaded edge-coupled phase shifters, such as in [4], is
large enough to produce significant reactive discontinuities in
practical transverse electromagnetic line geometries, adversely
affecting the phase accuracy of the devices. Tresselt described
a design that could alleviate that effect by employing coupling,
which is continuously tapered through the length of the device.
He designed and constructed a phase shifter for broadband applications. However, the results indicated that the interconnecting strap parasitics, physical transitions, and etching tolerances (due to a tight coupling requirement), limited the upper frequency band and could only be partially compensated for, thus limiting the device application to around 8.5 GHz. In [10], two coupled, designed according to the tapered coupled method, were connected in tandem to form a differential phase shifter. The design required the use of a nine-section structure for the transmission lines in addition to several impedance transformers. The results presented in the paper show an insertion loss higher than 2 dB, and a phase error that is $\pm 5^\circ$ and $\pm 10^\circ$ in the 45° and 90° phase shifters, respectively.

In another approach to improve performance of the edge-coupled phase shifters, Taylor and Prigel [11] used a wiggling technique to design a broadband phase shifter. The wiggled edged coupled microstrip lines were used as a means of slowing the odd-mode microwave-energy propagation velocity to equal the even-mode propagation velocity and achieve broadband operation. The results in [11] show narrowband characteristics and suggest fabrication difficulties because of the very narrow space required between the coupled lines to accomplish a good performance.

In addition to the coupled lines structures, some other methods have been used to build planar phase shifters. Ahn and Wolff [12] presented several asymmetric ring-hybrid phase shifters. Each consists of a ring hybrid and reflecting terminations. The measured and simulated results in [12] indicated that the proposed design does not have the broadband characteristics of the edge-coupled structures.

With the rapid growth of microwave integrated-circuit technology, the switched phase shifters have been largely investigated [13]–[15]. The main target behind this type of phase shifter is to get a wide range of phase variation using the same device. In [13], a switching network was combined with a Schiffman phase shifter to build 180°-bit phase shifter. The network is composed of a half-wavelength coupled line, and parallel eighth wavelength open and short stubs, which are shunted at the edge points of a coupled line. The measured performance of the design shows a high phase deviation ($\pm 10^\circ$) over the band, i.e., 1.5–4.5 GHz. In [14], switching diodes were used to convert a microstrip line to a rectangular waveguide, whereas in [15], a branch line coupler controlled by a varactor diode was used. The common features of the switched phase shifters are a high insertion loss and a narrow bandwidth. In another method, a 3-D electromagnetic-bandgap woodpile was used to design a phase shifter, which is equivalent to the switched type [16]. It suffers from the same limitations of the switched type.

Three papers have recently appeared, which suggest modifications on the previously designed phase shifters to improve their performance [17]–[19]. A compensation technique was introduced in [17] and [18] to improve performance of the Schiffman phase shifter and the multistage design proposed by Shelton and Mosko [4]. Five compensating capacitors were used to improve the return-loss performance of the two circuits. The measured results indicated an improvement in the return-loss performance across the L-band. However, the use of the compensation technique in [17] did not increase the useful phase-stable bandwidth; actually it resulted in a 2–3-GHz bandwidth. The use of the compensation technique in [18] for the multistage phase shifter increased the insertion loss of the device. No results were given in [18] to show the effect of the additional compensating elements on the phase performance of the phase shifter.

The latest modification to the Schiffman phase shifter included altering the ground plane underneath the coupled lines [19]. The ground plane under the coupled lines was removed; meanwhile an additional isolated rectangular conductor was placed under the coupled lines to act as a capacitor. This modification enabled the designer to build a compact phase shifter (dimension $\approx 3$ cm $\times$ 4 cm). The measured results for this device show that it covers a 2:1 frequency band with better than 12-dB return loss and a moderate phase imbalance ($\pm 5^\circ$). It still has the problem of a need for a narrow gap between the coupled lines, especially at the high-frequency range of the ultra-wideband (UWB). Moreover, the bandwidth coverage of the device indicates that it cannot be used for UWB application where the band extends from 3.1 to 10.6 GHz (3.42 : 1 band).

For the edge-coupled structure, which was used by most of the previously mentioned designs, the coupling factor is largely dependent on the gap between the two coupled lines and the dielectric constant of the substrate. Therefore, the edge-coupled phase shifters are very difficult to be fabricated using microstrip lines on printed circuit board technology for UWB applications, as the gap between the coupled lines must be very narrow to obtain a tight coupling.

In a previous study by the author to build directional couplers [20], it was noticed that broadside coupling can achieve UWB characteristics without fabrication difficulties compared with edge coupling. The work presented here adopts the broadside coupling strategy using an elliptical coupled structure in the design of UWB phase shifters. There are many challenges that are addressed in this paper: how to build a two-port broadside-coupled phase shifter with minimum insertion loss and maximum return loss across the 3.1–10.6-GHz band; how to derive a theoretical model, which shows the relation between the phase shift and the coupling factor; and how to achieve a constant phase shift across the UWB.

The proposed method in this paper exploits broadside coupling between top and bottom elliptical microstrip patches via an elliptical slot located in the mid layer, which forms the ground plane. Variations of the phase shift, return loss, and insertion loss of the device with the coupling factor are calculated using a simple theoretical model. The model shows that it is possible to design high-performance UWB phase shifters for the 25°–48° range using the suggested structure. The proposed method is used to design 30° and 45° phase shifters that have compact size. The simulated and measured results show that the designed phase shifters achieve better than $\pm 3^\circ$ phase stability, less than 1-dB insertion loss, and better than 10-dB return loss across the UWB, i.e., 3.1–10.6 GHz. In addition to that, the presented device has a simple structure, which can be easily manufactured.

II. Analysis

The analysis used in this paper follows the conventional approach adopted for the coupled microstrip lines [21]–[24]. The phase shifter is considered as a four-port device with two of its
where $l$ is the physical length of the coupled structure and $\xi_{f}$ is the effective phase constant in the medium of the coupled structure. For the structure under investigation, it is possible to show that

$$\bar{\xi}_{f} = \frac{\xi_{e} + \xi_{o}}{2} = \sqrt{\varepsilon_{r}} \phi \times \sqrt{\frac{2}{\lambda}}$$

where $\xi_{e}$ and $\xi_{o}$ are the phase constants for the even and odd modes, respectively, $\lambda$ is the free-space wavelength, and $\varepsilon_{r}$ is the dielectric constant of the substrate.

Assuming that the output port (Port 2) is perfectly matched, then the incident ($\times$ reflected) signals at ports 3 and 4 are

$$l_{3} = \frac{j \bar{\xi}_{f} \sin(\chi_{e} l_{f})}{\sqrt{1 - \left(\frac{l_{f}}{2}\right)^{2} \cos(\chi_{e} l_{f}) + j \sin(\chi_{e} l_{f})}}\quad \text{(4)}$$

$$l_{4} = \frac{\sqrt{1 - \left(\frac{l_{f}}{2}\right)^{2} \cos(\chi_{o} l_{f}) + j \sin(\chi_{o} l_{f})}}{\sqrt{1 - \left(\frac{l_{f}}{2}\right)^{2} \cos(\chi_{o} l_{f}) + j \sin(\chi_{o} l_{f})}} \quad \text{(5)}$$

As ports 3 and 4 are terminated in an open circuit, then the reflection coefficient at those ports is equal to 1. Therefore, $l_{3} = l_{4}$ and $\alpha_{3} = \alpha_{4}$. Using this conclusion in (1)–(5),

$$\bar{\xi}_{31} = \frac{1 - \left(\frac{l_{f}}{2}\right)^{2} \left(1 + \sin^{2}(\chi_{e} l_{f})\right)}{\sqrt{1 - \left(\frac{l_{f}}{2}\right)^{2} \cos(\chi_{e} l_{f}) + j \sin(\chi_{e} l_{f})}} \quad \text{(6)}$$

$$\bar{\xi}_{21} = \frac{\sqrt{1 - \left(\frac{l_{f}}{2}\right)^{2} \cos(\chi_{o} l_{f}) + j \sin(\chi_{o} l_{f})}}{\sqrt{1 - \left(\frac{l_{f}}{2}\right)^{2} \cos(\chi_{o} l_{f}) + j \sin(\chi_{o} l_{f})}} \quad \text{(7)}$$

where $\bar{\xi}_{31} = l_{1} / \alpha_{1}$ is the return loss of the input port, $\bar{\xi}_{21} = l_{2} / \alpha_{2}$ is the insertion loss from the input to the output port.

Phase shift of the output signal compared to the input signal can be found from (7) as follows:

$$\bar{X}_{n} = \frac{\bar{X}_{f} - 2 \arctan \left[ \frac{\sin(\chi_{e} l_{f})}{\sqrt{1 - \left(\frac{l_{f}}{2}\right)^{2} \cos(\chi_{e} l_{f}) + j \sin(\chi_{e} l_{f})}} \right] \right] \quad \text{(8)}$$

To find the differential phase shift, which can be obtained using the proposed structure, a comparison should be made with a reference transmission line. A 50-$\Omega$ microstrip line is assumed as a reference in this paper. The phase shift caused by a section of microstrip line of physical length $l_{n}$, is

$$\bar{X}_{n} = - \bar{\xi}_{n} \alpha_{n} = - \left(\frac{l_{n}}{\alpha_{n}}\right) \times \left(\frac{j}{\sqrt{\varepsilon_{r}}} \phi \times \sqrt{\frac{2}{\lambda}}\right) \quad \text{(9)}$$

where $\bar{\xi}_{n}$ and $\alpha_{n}$ are the phase constant and effective dielectric constant of the microstrip transmission line, respectively. The parameter $\varepsilon_{r}$ can be calculated using the well-known formulas as in [25].

The differential phase shift ($\Delta \bar{X}_{P}$) from (8) and (9) is

$$\Delta \bar{X}_{P} = \bar{X}_{f} - 2 \arctan \left[ \frac{\sin(\chi_{e} l_{f})}{\sqrt{1 - \left(\frac{l_{f}}{2}\right)^{2} \cos(\chi_{e} l_{f}) + j \sin(\chi_{e} l_{f})}} \right] + \bar{\xi}_{n} \alpha_{n} \bar{\xi}_{f} l_{n} \quad \text{(10)}$$

Variation of the calculated $\Delta \bar{X}_{P}$, using (10), for different values of the coupling ($\chi_{e} l_{f}$) and coupling factor ($\phi$) is shown in Fig. 3. The Rogers RO4003C (with $\varepsilon_{r} = \gamma_{c} \phi$, thickness = 1.5625 mm, and tangent loss = $\tan(\phi)\phi$) was assumed as the substrate.
In Fig. 3, the physical length of the microstrip transmission line \( l_{\text{p}} \), is optimized to obtain a minimum deviation in the differential phase shift for each value of the coupling. It was found that \( l_{\text{p}} \) should be around \( 2L \), which is the total physical length of the top and bottom coupled patches.

Fig. 3 indicates an odd symmetry of \( \Delta \Phi \) around the length \( \lambda_{\text{g}}/2 \). The results in Fig. 3 also reveal that it is possible to design a phase shifter with wide range of phase shift by varying value of the coupling factor \( \zeta \). The estimated phase range using (10) extends from 0° to 90° for \( \zeta \) from 1 down to 0. There are still two parameters to be checked before judging performance of the device: the return loss and insertion loss.

According to the results shown in Fig. 3, there is an inverse relation between \( \Delta \Phi \) and \( \zeta \). This conclusion is verified by plotting in Fig. 4 the average value of \( \Delta \Phi \) and the phase deviation (with respect to the average value) with \( \zeta \) using the data of Fig. 3. It is also obvious from this figure that the maximum phase deviation around the nominal value of \( \Delta \Phi \) has a direct proportional relationship with \( \zeta \).

Designing a high-performance phase shifter not only requires phase stability with the least deviation around the nominal value across the required bandwidth, but it also requires that the device should have a low insertion loss and a high return loss across that band. Variation of the return loss and insertion loss with the coupling length are shown in Figs. 5 and 6 for different values of the coupling factor by using the absolute value of the parameters in (6) and (7).

It is clear from Figs. 5 and 6 that there is an even symmetry of the return loss and insertion loss around the point \( \lambda_{\text{g}}/2 \). Hence, to achieve the best performance (low insertion loss and high return loss) over a broad band, \( \lambda_{\text{g}}/2 \) should be equal to 90° at the center frequency of operation.

Referring to Figs. 5 and 6, it is important to make sure which values of \( \zeta \) give an acceptable performance over the UWB from 3.1 to 10.6 GHz. As mentioned earlier, the length of the coupled structure must be 90° at the center frequency, which is \( (\lambda_{\text{g}} + l \zeta)/2 \) from 3.1 to 10.6 GHz. Assuming the physical length...
of the coupled structure $l$ is constant, then the coupling length $k_{\ell} l$ is equivalent to $\frac{\pi}{2} \times \frac{3.1}{6.5} = 4.17\degree$ at 3.1 GHz and $\frac{\pi}{2} \times \frac{10.6}{6.5} = 13.8\degree$ at 10.6 GHz. According to Figs. 5 and 6, the return loss is more than 10 dB and the insertion loss is less than 0.5 dB across the whole UWB (3.1–10.6 GHz) when $1.72 < \zeta < 10.82$.

From Fig. 4, it seems that it is possible to design high-performance 25°–48° phase shifters that cover the 3.1–10.6-GHz band using the proposed model with $1.72 < \zeta < 10.82$. A multistage phase shifter can be designed to achieve a higher range of differential phase shift when required. If, in some applications, the required bandwidth is less than the one used in this paper, other ranges of phase shifts are possible with the presented method using only one stage.

### III. Design

In order to establish the validity of the proposed method, 30° and 45° phase shifters were designed using the following steps. From Fig. 4, the coupling should be 0.81 and 0.73 for the 30° and 45° phase shifter, respectively. The return loss, from Fig. 5, is higher than 12 and 10 dB for the 30° and 45° phase shifter, respectively, across the 3.1–10.6-GHz band. The insertion loss is less than 0.4 and 0.5 dB for the 30° and 45° phase shifter, respectively, across the same band (see Fig. 6).

Depending on the value of the coupling, the even ($Z_{oe}$) and odd ($Z_{oo}$) mode characteristic impedances for the coupled patches are calculated using the following equations:

$$Z_{oe} = \frac{Z_0 \sqrt{1 + \zeta}}{\sqrt{1 - \zeta}}, \quad Z_{oo} = \frac{Z_0 \sqrt{1 - \zeta}}{\sqrt{1 + \zeta}}$$

where $Z_0 = \frac{\varepsilon_r}{\sqrt{2}}$ is the characteristic impedance of the microstrip ports of the coupler. Using $\zeta = 1.81$ for the 30° phase shifter, the impedances $Z_{oe}$ and $Z_{oo}$ can be found to be 154.3 and 16.2\i ohms, respectively. If the device is designed to have 45° phase shift, $Z_{oe}$ and $Z_{oo}$ can be calculated to be 126.6 and 19.8\i ohms, respectively. To determine dimension of the coupled region, which gives these impedance values, it is possible to use the following equations [26]:

$$Z_{oe} = \frac{\varepsilon_r \pi \sqrt{K'(l_1)}}{\sqrt{\varepsilon_r \pi \sqrt{K'(l_2)}}}, \quad Z_{oo} = \frac{\varepsilon_r \pi \sqrt{K'(l_2)}}{\sqrt{\varepsilon_r \pi \sqrt{K'(l_1)}}}$$

where $\varepsilon_r$ is the dielectric constant of the substrate, $K'(l_1)$ is the first kind elliptical integral, and $K'(l_2) = K'(\sqrt{1 - l_2^2})$. The parameters $l_1$ and $l_2$ are used to find the major diameters of the elliptical coupled microstrip at the top and bottom layers ($T_{ap}$) and slot ($T_{as}$) at the mid layer according to the following equations [20]:

$$l_1 = \frac{\sinh^2{\left(\pi^2 T_{ap}/(1+i)\right)}}{\sqrt{\sinh^2{\left(\pi^2 T_{ap}/(1+i)\right)} + \cosh^2{\left(\pi^2 T_{ap}/(1+i)\right)}}}$$

$$l_2 = 1.0 \sinh{\left(\pi^2 T_{ap}/(1+i)\right)}$$

Physical length ($l_2$) of the elliptical microstrip/slot coupled structures ($l$) must be chosen to be equal to a quarter of the effective wavelength at the center frequency of operation, i.e., at 6.85 GHz, as proven according to Figs. 3 and 4. It is to be noted that the coupling factor $\zeta$ was considered constant when calculating each set of results shown in Figs. 3, 5, and 6. However, the measured results for the directional couplers in [20], which used a broadside coupled structure similar to the one used in this paper, show that the coupling factor $\zeta$ tends to be lower at the two ends of the frequency band compared with its value at the center frequency. If this effect is included into the results shown in Fig. 4, the designed device is going to have a higher average phase shift at the two ends across with its value at the center frequency. On the other hand, Fig. 3 shows that, for a certain value of the coupling, $\Delta \phi$ is larger than the average value at the lower frequency band and smaller at the upper frequency band. The combination of these two factors results in a worse performance at the lower end and a better performance at the upper end of the frequency band. Therefore, it is better to design the phase shifter with a length that is larger than 90° at the center frequency.

A comparative study of the mode of variation of $\Delta \phi$ (from Fig. 3) and measured $\zeta$ (from [20]) with frequency indicated that the widest bandwidth can be achieved when the coupling length $8_2 l$ is greater than 111$\pi$ F at a center frequency of 6.85 GHz.

The last step in the design procedure is to find width ($w$) of the reference line and the microstrip lines that connect the phase shifter to the 50-ohm input/output ports. This can be achieved using the standard microstrip design equations [25].

Dimension of the phase shifters calculated using the proposed method are shown in Table I. The phase shifters were assumed to use a Rogers RO4003C as a substrate. Table I also shows the final dimension after fine tuning using the optimization capability of the full electromagnetic software package Ansoft HFSSv10. There is a little difference (less than 5%) between the calculated and optimized values. This gives a high credibility to accuracy of the proposed method.

### IV. Results and Discussions

To prove the validity of the presented design method, the 30° and 45° phase shifters designed in Section III and aimed for the operation in the 3.1–10.6-GHz frequency band were manufactured and tested. A Rogers RO4003C, with 17-\mu m thick conductive coating, is selected for the devices development. A photograph for one of the manufactured phase shifters is shown in Fig. 7. Dimension of the phase shifter alone (excluding the reference transmission line) is 2.5 cm × 2 cm. This reveals compactness of the proposed phase shifter.

It is to be noted that, in the manufactured devices, the coupled structure of physical length $l$ and the reference transmission line of physical length $l_2$ are connected to the input/output subminiature A (SMA) connectors using the same additional length of microstrip transmission lines.
Fig. 7. One of the manufactured phase shifters. (a) Top layer. (b) Bottom layer. The upper part of (a) and (b) is the phase shifter, whereas the reference transmission line is shown at the lower part of (a).

Fig. 8. Simulated and measured differential phase shift for the two developed phase shifters.

The differential phase shift $\Delta \phi$, return loss, and insertion loss of the designed devices were first verified using the Ansoft HFSSv10 software and then measured using a vector network analyzer. The simulated and measured $\Delta \phi$ of the designed $30^\circ$ and $45^\circ$ phase shifters are shown in Fig. 8. It is clear that the designed phase shifter features UWB characteristics. The value of $\Delta \phi$ is $30^\circ \pm 3^\circ$ in the simulations and $30^\circ \pm 2.5^\circ$ according to the measured results for the $30^\circ$ phase shifter across the 3.1–10.6-GHz band. For the $45^\circ$ phase shifter, $\Delta \phi$ is $45^\circ \pm 3^\circ$ in the simulation and $45^\circ \pm 2.3^\circ$ according to the measured results across the same UWB. Fig. 8 indicated that the measured results are close to the simulated results, and both of them are in good agreement with the theoretical prediction shown in Fig. 4, which gives an estimation of $30^\circ \pm 2.7^\circ$ and $45^\circ \pm 2.4^\circ$ for the differential phase shift for the two phase shifters. The general shape of differential phase variation with frequency is also in good agreement with results of the theoretical analysis shown in Fig. 3.

The combined effect of using a coupling length equal to 110 (instead of 90) and the nonconstant value of $\gamma$ on $\Delta \phi$ can be seen in Fig. 8. The measured value of $\Delta \phi$ at the lower frequency band is almost equal to its value at the higher frequency band for the $45^\circ$ phase shifter. In the case of the $30^\circ$ phase shifter, $\Delta \phi$ is almost constant during most of the frequency band.

The simulated and measured return loss for the $30^\circ$ phase shifter is better than 12 dB according to the simulations and better than 10 dB in the measured results across the whole UWB (see Fig. 9). It is always better than 15 dB in the 3.7–11-GHz band. Concerning the $45^\circ$ phase shifter, the return loss is better than 10 dB in the 3.3–10.6-GHz band, as shown in Fig. 9. The return loss of this phase shifter is always better than 18 dB in the 4.2–10.2-GHz band. There is a good agreement between the measured and simulated results shown in Fig. 9 and the theoretical estimation shown in Fig. 5, except a little discrepancy at the lower part of the frequency band, i.e., around 3 GHz. This discrepancy can be justified by the combined effect of using a longer coupled structure and the nonconstant value of $\gamma$.

In Fig. 10, the simulated and measured insertion loss for the $30^\circ$ and $45^\circ$ phase shifters is shown. The simulated insertion loss for the $30^\circ$ phase shifter is better than 0.6 dB, whereas it
is better than 0.85 dB according to the measured results in the 3.1–10.6 GHz band. Concerning the 45° phase shifter, the simulated and measured insertion loss is better than 0.8 and 1 dB, respectively, across the 3.1–10.6 GHz band, as shown in Fig. 10.

There is a little difference between the measured and simulated results shown in Fig. 10. The measured insertion loss is more than the simulated value by approximately 0.15 dB, on average, for the two phase shifters. This additional insertion loss comes from the two SMA connectors, which were used in the measurements, but not included in the simulations. According to the data sheet of the used connectors, their insertion loss increases with frequency and becomes more than 0.1 dB per connector after 6 GHz.

V. CONCLUSION

Simple and clear guidelines have been presented to design compact planar phase shifters with UWB characteristics. The proposed method exploits broadside coupling between top and bottom elliptical microstrip patches via an elliptical slot located in the mid layer, which forms the ground plane. A theoretical model has been used to analyze performance of the proposed devices. The model has shown that it is possible to design high-performance UWB phase shifters for the 25°–45° range using the proposed structure. The design method has been used to design 30° and 45° phase shifters, which have compact size, i.e., 2.5 cm × 2 cm. The simulated and measured results have shown that the designed phase shifters have better than ±3° phase stability, less than 1-dB insertion loss, and better than 10-dB return loss across the UWB from 3.1 to 10.6 GHz.

The UWB behavior, compactness, and easy of fabrication of the presented phase shifters should attract considerable interest from designers of wireless systems, in general, and UWB systems, in particular.

The multilayer broadside-coupled configuration of the proposed device is especially suitable for implementation in modern multilayer structures such as the laminated multichip modules (MCMs-L) and low temperature co-fired ceramics (LTCC). In such structures, broadside coupling is much preferred from a reproducibility and loss perspective.

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Amin M. Abbosh was born in Mosul, Iraq. He received the M.Sc. degree in communication systems and Ph.D. degree in microwave engineering from Mosul University, Mosul, Iraq, in 1991 and 1996, respectively.

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efficiency according to temperature are large. But, a GaN HEMT PA is less sensitive to temperature than a Si LDMOS PA, as shown in Figure 8(b). The small ACLR degradation due to temperature is easily restored by the adaptive control of the vector modulator in the PA.

Figure 9 shows the measured characteristics of Si LDMOS and GaN HEMT PAs according to the variation of operating frequency for a 4-carrier WCDMA signal at a Pout of 34 dBm. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
Phase shifters are important components for many applications, such as phase modulators, phased array antennas, radars, and measurement systems. In these and many other applications, and to get wideband performance, phase shifters are usually realized in planar technology using the stripline or microstrip lines. The factors that should be considered when designing a phase shifter are bandwidth, differential phase shift, phase errors, insertion losses, complexity, and size.

To achieve the broadband operation of the phase shifters, the approach of the coupled transmission lines is usually employed. One of the earliest designs that used the coupled lines method to construct broadband phase shifters is Schiffman differential phase shifter [1]. Although it can achieve 90° differential phase shift over a broadband, the measured results of Schiffman’s phase shifter indicate a high phase ripple (± 10°). To design an edge-coupled phase shifter with an acceptable phase ripple, some authors proposed the use of the tandem configuration [2] and the cascaded multiple coupled sections [3–5]. However, tandem configuration requires wire crossovers, which is inconvenient from the manufacturing point of view, whereas to achieve a broadband performance in the cascaded connection, an extremely tight coupling is still required which may not be realizable in a given practical configuration.

A compact version of the Schiffman phase shifter was introduced in [6]. Although the proposed design uses a smaller area and it is a cost effective one compared with the original design, it has a narrow bandwidth. In another approach to improve the performance of the edge-coupled phase shifters, Taylor and Prigel [7] used a wiggling technique to design a broadband phase shifter. The design presented in [7] shows narrowband characteristics and suggests fabrication difficulties because of the very narrow space required between the coupled lines to accomplish a good performance.

Recently, a compensation technique has been proposed as a possible solution to improve the performance of the previously designed phase shifters [8]. However, the use of the proposed compensation technique did not increase the useful phase-stable bandwidth; actually it resulted in a higher insertion loss.

The reflection type phase shifter has also been investigated [9]. However, it has the disadvantage of narrow bandwidth, large phase error, and large size due to the design’s dependency on wavelength. In a recent design [10], a slow wave microstrip line has been used to make the reflection type phase shifter compact in size. However, it still has a narrowband performance.

In this article, a quadrature directional coupler and the Wilkins combiner are connected in series to form a fixed phase shifter with broadband performance. The quadrature coupler, which is used in the presented design, is an edge-coupled line with a slotted ground plane to relax the required gap between the coupled lines and make the manufacturing process easy.

2. ANALYSIS
A diagram of the proposed phase shifter is shown in Figure 1. It is composed of two devices: a 3-dB quadrature directional coupler and the Wilkinson combiner. The two components of the phase shifter are integrated in one printed circuit board.

The operation of the fixed phase shifter can be explained as follows. Assuming that the input signal to the device is \(a_0\) then the direct and coupled output from the quadrature coupler \(a_1\) and \(a_2\), respectively, are given as:

\[
a_1 = \frac{a_0 e^{-j\beta_1 l_1}}{\sqrt{2}}
\]

\[
a_2 = \frac{a_0 e^{-j(\beta_1 + \pi/2)l_2}}{\sqrt{2}}
\]

where \(l_1\) is length of the quadrature coupler. Note that the fourth port of the coupler, i.e. the isolated port, is connected to a matched load.

The two signals \(a_1\) and \(a_2\), which are the outputs from the direct and coupled ports of the coupler, are the inputs to the second stage of the phase shifter, i.e. the combiner. The output from the combiner \(a_c\), which is also the output from the phase shifter, is equal to:

\[
a_c = \frac{a_0 e^{-j(\beta_1 + \pi/2)l_2}}{\sqrt{2}} [1 + e^{-j(\pi/2 l_2)}] = a_0 e^{-j(\beta_1 + \pi/4)l_2}
\]

where \(l_2\) is the length of the combiner.

To find the differential phase shift due to the use of the proposed device, it is possible to compare between the phase shift owing to the device and the phase shift due to the use of a reference transmission line with the same signal input and the same length, which should be \((l_1 + l_2)\). The signal output from the reference line, which is a 50-Ω microstrip transmission line in this case, is equal to:

\[
a_r = a_0 e^{-j(\beta_1 + \pi/2)l_2}
\]

Comparing (3) with (4), the differential phase shift \(\Delta \Phi\) between the output signal from the proposed device and the output from the reference transmission line is equal to \(\pi/4\) (or 45°).

3. DESIGN
Configuration of the proposed 45° phase shifter is shown in Figure 2. It consists of two parts connected in series: the first part is a backward quadrature directional coupler, whereas the second part is the Wilkinson power combiner. The quadrature coupler is designed using a double sided printed circuit board. The top layer contains the two coupled microstrip lines with a length equal to quarter of the effective wavelength calculated at the centre frequency of operation, whereas the ground plane is located at the bottom layer, see Figure 2. There is a rectangular slot made at the
ground plane underneath the coupled lines. The slot at the ground plane was used to relax the requirement for a narrow spacing between the edge-coupled lines when the tight coupling of 3 dB is needed, as for the device presented in this article. The coupled lines of the coupler are twisted in the proposed structure to make length of the paths for the two signals ($a_1$ and $a_2$ in Fig. 1) equal. The isolated port of the coupler is terminated in a matched load, as shown in Figure 2.

The coupler can be fully analyzed and its initial dimension can be estimated using the even- and odd-mode analysis [11], the conformal mapping [12] and the image technique [13] following the steps which are explained briefly hereafter. If the input/output impedance of the coupler is $Z_0/\sqrt{50}$ and the coupling factor is 3 dB then values of the required even- and odd-mode impedances of the coupled-lines can be found to be 120.5/\sqrt{50}$ and 20.7/\sqrt{50}$, respectively, by using the formulas in [11]. Assuming a quasi transverse electromagnetic propagation, the relation between the calculated even- and odd-mode impedances of the coupled-lines and its dimensions can be found using the conformal mapping technique [12] and the image theory [13]. From which, the initial values of the design parameters can be estimated. The optimization capability of the software Ansoft HFSSv10 can then be used to obtain the final dimensions of the coupler.

The second part of the device is the Wilkinson combiner [11]. The two output signals from the quadrature coupler are connected to the two inputs of the combiner using a 50-\Omega microstrip lines with width equal to $w_f$ till the location of a junction, where the 100-\Omega resistance is connected, see Figure 2. After which, two transmission lines with length equal to quarter wavelength calculated at the centre frequency of operation (6 GHz), and a width $w_{\text{mic}}$ which gives a characteristic impedance equal to $\sqrt{50}$ = 70.7\Omega, are used. The final stage of the combiner, and the phase shifter, is a 50-\Omega microstrip line. Width of the microstrip lines required to build the combiner can be calculated using the design equations in [11].

4. RESULTS

The validity of the presented device was tested by designing a 45° phase shifter aimed for the operation across the C-band (4–8 GHz). Rogers RO4003C (with dielectric constant = 3.38, thickness = 0.508 mm, loss tangent = 0.0027) was selected as a substrate for the device. The software HFSSv10 was used to optimize values of the design parameters. The optimization was aimed to obtain the highest possible phase stability within the C-band. Assuming that the centre frequency of operation is 6 GHz (for the C-band), the optimized dimensions of the designed phase shifter ($w_c$, $w_s$, $w_f$, $s$, $l$, $w_{\text{mic}}$, and $l_{\text{mic}}$) are equal to 1, 7.2, 1.1, 0.13, 8.2, 0.6, and 6.7 mm, respectively. The overall dimension of the phase shifter is 30 mm x 30 mm indicating a compact size.

The differential phase shift ($\Delta \Phi$), return loss, and insertion loss of the designed device were verified using the software Ansoft HFSSv10, whereas the measurements were done using a vector network analyzer. The simulated and measured $\Delta \Phi$ of the designed 45° phase shifter is shown in Figure 3. It is clear that the designed phase shifter features broadband characteristics with the differential phase shift equals to $45° \pm 5°$ across the C-band (4–8 GHz).

In addition to the phase stability, a high performance phase shifter should have a low insertion loss from the input to the output
The shifter has a compact size with dimensions of 30 mm.

In high data rate systems, the designed phase delay enables its use in narrow pulse transmission. The designed device has also shown a flat group 1-dB insertion loss and better than 12-dB return loss at the centre of the C-band. The designed device has almost a constant group delay (less than 0.05-ns fluctuation around the mean value). The phase shifter presented in this article has almost a constant group delay (less than ±0.05-ns peak fluctuation) across the C-band as depicted in the simulated results in Figure 5.

There is another parameter that can be used to assess the performance of the phase shifter: it is the group delay [11]. To handle a narrow pulse operation or a high data rate transmission/reception, the group delay of the device should show a low fluctuation around the mean value. The phase shifter presented in this article has almost a constant group delay (less than ±0.05-ns peak fluctuation) across the C-band as depicted in the simulated results in Figure 5.

5. CONCLUSION

The design of a broadband fixed phase shifter for C-band applications has been presented. The proposed device is composed of two parts: a quadrature 3-dB directional coupler and the Wilkinson combiner. The directional coupler utilized in the design is based on the edge-coupled microstrip lines. A slotted ground plane was used underneath the coupled lines to relax the required gap between the coupled lines and make the manufacturing process easy. The phase shifter presented in this article has shown a 45° ± 5° fixed differential phase shift across the band 4–8 GHz with less than 1-dB insertion loss and better than 12-dB return loss at the centre of the C-band. The designed device has also shown a flat group delay, which enables its use in a narrow pulse transmission/reception and in the high data rate systems. The designed phase shifter has a compact size with dimensions of 30 mm × 30 mm.

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Broadband Fixed Phase Shifters

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Abstract—A method to design planar and compact phase shifters with broadband characteristics is presented. It utilizes broadside-coupled microstrip-coplanar waveguide, and thus the proposed devices can be fabricated using the simple and cheap double-side printed circuit boards. The method is used to design 60° and 90° phase shifters. The simulated and measured results show that the developed phase shifters achieve 3 to 11 GHz bandwidth with low phase instability (±2°), very low insertion loss (0.4 dB), high return loss (15 dB), and a compact size (1.5 cm × 2 cm).

Index Terms—Coupled transmission lines, electromagnetic coupling, phase Shifters, planar transmission lines.

I. INTRODUCTION

Phase shifters are key devices in many microwave systems, such as intelligent antennas, microwave instruments, modulators, to name a few.

The conventional approach to design planar phase shifters is to use the Schiffman differential phase shifter or one of its variations that rely on the edge-coupled transmission lines [1]. However, a very narrow gap between the edge-coupled lines is needed for a broadband performance.

In an important development, multilayer broadside-coupled structures were utilized to build ultra-wideband (UWB) phase shifters with excellent performance [2]. This multilayer phase shifter has recently been utilized to build UWB Butler matrix for switched beam antenna array that operates across the range 3.1 to 10.6 GHz [3]. However, the multilayer configuration is not the preferred option for some applications which require the use of the simple and low-cost printed circuit board (PCB) technology. Moreover, the multilayer structure could pose some manufacturing challenges as any error in the alignment of the different layers may cause a significant degradation in the performance.

In recent developments, reflective-, active-, liquid crystal polymer-, metamaterials-, and substrate integrated waveguide-based phase shifters are proposed [4]–[10]. However, all of those types have either a limited relative bandwidth (10%–40%) or a high insertion loss (2 to 5 dB), in addition to their complicated and costly manufacturing process.

In this letter, a broadside-coupled microstrip-coplanar waveguide (CPW) structure is utilized to develop broadband phase shifters. The proposed approach enables the use of PCB to develop phase shifters with low phase variation (2°), very low insertion loss (2°), and high return loss (15 dB).

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If the device is designed properly such that \( \phi_{11} = 0 \), the differential phase shift \( \Delta \phi \) of the output signal compared to a normal microstrip transmission line of length \( l_4 \), and phase constant \( \kappa_0 \) can be found from (1a)–(1c)

\[
\Delta \phi = 1 + \phi - \pi \text{artanh} \left( \frac{a_1 \phi}{\sqrt{1 - \phi^2}} \right) + \kappa_0 l_4 \tag{2}
\]

The main reason behind the possibility of achieving a constant \( \Delta \phi \) as given in (2) across the UWB is the constant \( \kappa \) of the utilized structure across that band. The effect of the variation in \( \kappa \) with frequency can be compensated by the reverse effect of \( \kappa_0 l_4 \). This step requires the optimization of \( l_4 \) for a minimum deviation in \( \Delta \phi \) across the UWB.

Following the analysis presented in [2], it is possible to show that, for a UWB performance, the proposed structure can achieve a phase shift across the range from 30° to 90°. The coupling factor \( \kappa \) is the parameter that can be used to control value of the differential phase shift. After calculating the value of \( \kappa \) needed to achieve the required phase shift from (2), the physical dimensions for the structure are to be found. Thus, the relation between the coupling factor \( \kappa \) and the physical dimensions of each section of the device shown in Fig. 1 needs to be derived.

The structure depicted in Fig. 1 can be analyzed using the \( \phi \) - and \( \pi \)-mode approach. For the \( \phi \) -mode, the two layers are excited in-phase, whereas in the \( \pi \)-mode, the top and bottom layers are out-of-phase with respect to the ground.

Assuming a quasi transverse electromagnetic propagation, the electrical characteristics of the coupled lines can be completely determined from the effective capacitances per unit length of the lines and the phase velocity on the lines [11]. Therefore, the structures shown in Fig. 2 can be used to analyze the proposed device.

For each of the two modes of propagation, the capacitance for each of the two coupled lines can be determined from Fig. 2. The \( \phi \) -mode capacitance for the microstrip \( \phi \) and the \( \pi \) PW \( \pi \) are equal to

\[
\phi_{\text{MS}} = \phi_{\text{PW}} \pi \quad \phi_{\text{CW}} = \phi_{\text{PW}} \pi
\]

The \( \pi \)-mode capacitance for the microstrip \( \pi \) and the \( \pi \) PW \( \pi \) are equal to

\[
\phi_{\text{MS}} = \phi_{\text{PW}} \pi + 2 \phi_{\text{MS}} \pi \pi = \phi_{\text{PW}} \pi + 2 \phi_{\text{MS}} \pi
\]

Using the quasi-static approach with the help of the conformal mapping technique, the capacitances shown in Fig. 2 can be calculated as a function of the coupled structure’s dimension

\[
\phi_{\text{MS}} = 2 \varepsilon_{\text{MS}} \phi_{\text{PW}} \pi \quad \phi_{\text{MS}} = 2 \varepsilon_{\text{MS}} \phi_{\text{PW}} \pi
\]

The subscript \( i \) refers to the line (\( \pi \) for microstrip and \( \pi \) for CPW) and \( j \) refers to the mode (\( \pi \) for \( \phi \) -mode and \( \pi \) for \( \pi \)-mode), \( \varepsilon \) is velocity of light in free space, and \( \varepsilon_{\text{MS}} \) is the dielectric constant of the substrate.

The structure shown in Figs. 1 and 2 is asymmetrical. Therefore, the analysis approach for asymmetrical broadside-coupled lines is used [12]. Hence, it is possible to find the coupling factor between the top layer and the bottom layer (\( \kappa \)) using the following equation [12]:

\[
\kappa = \sqrt{\frac{Z_{\text{MS}} Z_{\text{PW}}}{Z_{\text{CW}} Z_{\text{CW}}}}
\]

The coupling factors as a function of the capacitances can be obtained by substituting from (3) and (5) into (6). The design (4)–(6) can now be used to find the initial dimensions \( \pi \) for the device.

Concerning lengths of the coupled structure \( l_1 \) and \( l_3 \), they are initially chosen to be equal to quarter of the effective wavelength at the centre of the passband (6.85 GHz). The length \( l_2 \) is less than \( l_3 \) by the value of the narrow slot needed in the ground plane to form the CPW.
III. RESULTS

To test the performance of the proposed device, 60° and 90° phase shifters were designed and fabricated using Rogers RT6010 (εr = 3.18, h = 0.508 mm) as the substrate. The proposed design method was used to calculate the initial dimensions of the devices, whereas the final dimensions were found using the optimization capability of the software HFSS. It was noted that the absolute difference between the initial and the optimized values is less than 12% reflecting the accuracy of the utilized theoretical analysis.

The optimized dimensions in (mm) are $P_{th} = 8.18$, $P_c = 4.38$, $l_1 = 3.02$, $l_2 = 3$, and $l_3 = 4.02$ for the 60° device, and $P_{th} = 8.18$, $P_c = 4.38$, $l_1 = 3.02$, $l_2 = 4.14$, and $l_3 = 4$ for the 90° device. The distance $d$ has no significant impact on the performance, and for a compact structure, it is chosen to be as small as possible; $d = 1.3$ mm for the 60° device, and 1.03 mm for the 90° device. Size of the devices excluding the reference line is $1.7 \times 1.2 \times 2$ cm. A photo of one of the developed devices is shown in Fig. 3.

The performance of the designed devices was verified using the software HFSS and then measured using a vector network analyzer. The simulated and measured $\Delta \phi$ of the 60°, and 90° phase shifters are shown in Fig. 4. The value of $\Delta \phi$ is $\Delta \phi \pm 1^\circ$ in the simulations and $\Delta \phi \pm 2^\circ$ according to the measured results for the 60° phase shifter across the band from 3.5 to 12 GHz. It is interesting to see from Fig. 4 that the achieved phase shift of this device is almost constant at 60° across the band from 4 GHz to more than 12 GHz. For the 90° phase shifter, $\Delta \phi$ is $\Delta \phi \pm 3^\circ$ for the band 3 to 11 GHz.

The S-parameters of the developed devices are shown in Fig. 5, which reveals a passband of 2.8 GHz to more than 12 GHz for the 60° phase shifter and 2.8 to 11 GHz for the 90° phase shifter assuming the 10 dB return loss as a reference. The insertion losses of the two phase shifters are less than 0.2 dB in the simulations and less than 0.4 dB in the measurements across those passbands. Fig. 5 also shows that the return losses of the two developed phase shifters are more than 15 dB across most of the passbands.

The performance of the developed devices is also tested when connected as a complete differential phase shifter using two single-pole double-throw switches. It was noted that the phase performance and the return loss are exactly as shown in Figs. 4 and 5, whereas the insertion losses increase by 0.45 dB due to the use of those switches as proven by their technical specifications.

IV. CONCLUSION

A broadband phase shifter that can achieve a constant differential phase has been presented. The device utilizes a broadside-coupled microstrip-CPW structure, which enables a simple and cheap manufacturing process using the printed circuit board’s technology. A complete design procedure has been presented for the proposed device that has a compact size of $1.5 \times 2.2 \times 0.14$ mm. The simulated and measured results of the developed 60°, and 90° phase shifters have shown 3 to 11 GHz bandwidth.

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ULTRA-WIDEBAND PHASE SHIFTER USING BROADSIDE-COUPLED COPLANAR WAVEGUIDE STRUCTURE

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ABSTRACT: A planar phase shifter with ultra-wideband performance is presented. The proposed device utilizes a single-section broadside-coupled coplanar waveguide structure that can be implemented using a double-sided printed circuit board. The conformal mapping technique is used to find the proper dimensions of the device. The simulated and measured results of a 45° phase shifter designed using the proposed structure show more than 110% fractional bandwidth with a maximum phase deviation of ±3.3° and maximum insertion loss of 0.8 dB. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 54:114–116, 2012; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.26451

Key words: differential phase shifter; ultra-wideband phase shifter; coplanar waveguide; coupled structure

1. INTRODUCTION

Phase shifters have numerous applications, such as phased arrays, microwave instrumentation and phase modulators, to name a few. With the recent introduction of the ultra-wideband (UWB) standard (3.1–10.6 GHz), there has been a huge interest in developing different types of microwave devices that have the required performance across the above-mentioned band. One of those devices is the phase shifter.

Microwave phase shifters should have a constant differential phase shift as compared with a conventional transmission line across the band of interest. They should also have a low insertion loss, and thus, a high return loss across that band. For UWB applications, those requirements should be maintained across the band from 3.1 to 10.6 GHz. One of the utilized approaches to develop a wideband phase shifter is to use T-shaped open stub loaded transmission line [1]. The measured results indicate an 83% bandwidth [1], which is not enough for UWB systems. In another approach that has been derived from Ref. 1, a combination of parallel stubs and ground slots are used to build a UWB phase shifter [2]. The dimensions of the utilized slots in the ground plane the stubs that load the main transmission line are obtained via trial-and-error. The experimental results show a relatively large phase deviation (±5°).

In a recent development, broadside-coupled microstrip/microstrip and microstrip/coplanar waveguide (CPW) structures have been proposed to design UWB phase shifters [3, 4]. Concerning the microstrip/microstrip configuration, multilayer structure with four dielectric and five conductive layers is needed, and thus, it is not suitable for the printed circuit board (PCB) technology. The microstrip/CPW configuration has recently been proposed [4] to eliminate the multilayer requirement of the previous broadside-coupled structure [3]. However, two coupled sections are needed to maintain the use of one type of transmission line for the easy integration with other microwave devices.

In this letter, a simple technique that can be implemented on a two-sided PCB using only one coupled section is proposed to design a UWB phase shifter. It utilizes the strong broadside electromagnetic coupling between a CPW at the top side and a similar structure at the bottom side of the substrate. A complete design method for the proposed device is presented. The simulated and measured results of a 45° phase shifter designed using the proposed method to validate its UWB performance.

2. PROPOSED PHASE SHIFTER

The geometry of the proposed device is shown in Figure 1. It consists of two CPW patches that are connected to the input and output ports facing each other at the top and bottom side of the substrate. The two patches are strongly coupled in the broadside direction.

To find the phase shift performance of the device, the differential phase shift is calculated by comparing the phase shift introduced by the proposed device with a 50-Ω reference CPW transmission line. In addition to the differential phase shift, the performance of the device is also defined by the return loss and insertion loss at the passband.

For phase shifters that utilize coupled structures, the coupling factor is the parameter that can be used to control the value of the differential phase shift. The relation between them is derived in Ref. 3. It is possible to show that to design, for example, 45° phase shifter, the coupling factor \( C_{14} \) should be 0.73.

Figure 1 Top (a) and bottom (b) view of the proposed device. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

Figure 2 Electric field lines in the (a) even-mode and (b) odd-mode. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]
The developed phase shifter with the reference transmission line. [Color figure can be viewed in the online issue, which is available at wileyonlinelibrary.com]

The structure depicted in Figure 1 can be analysed using the even–odd mode approach. For the even-mode, the top and bottom CPW are excited in-phase [Fig. 2(a)], whereas in the odd-mode, the two CPWs are excited out-of-phase [Fig. 2(b)]. To achieve a coupling factor equal to 0.73, the required even-mode impedance ($Z_{ee}$) and odd-mode impedance ($Z_{oo}$) for the coupled structure that has input/output port impedances of $Z_o$ can be calculated using the following relations

$$Z_{ee} = Z_o \sqrt{\frac{1 + c_i}{1 - c_i}} \quad (1)$$
$$Z_{oo} = Z_o \sqrt{\frac{1 - c_i}{1 + c_i}} \quad (2)$$

For 50-$\Omega$ port impedances, $Z_{ee}$ and $Z_{oo}$ are equal to 126.6 and 19.8 $\Omega$, respectively. Assuming a quasi transverse electromagnetic propagation, the electrical characteristics of the coupled lines can be determined using the effective capacitors per unit length of the lines and the phase velocity on the lines [3]. Therefore, the structures shown in Figure 2 can be used to calculate the effective total capacitors at the two modes. The even-mode capacitor per unit length is the sum of two components that represent the electric fields in the upper (or lower) free space and middle dielectric material. The expressions for these two capacitors can be found by mapping each of the regions first into a half-plane and then into the well-known parallel-plate configuration [5]. In the calculations, a magnetic wall is assumed along the line of symmetry passing at the middle of the substrate as shown in Figure 2(a). The total odd-mode capacitor per unit length is calculated in the same manner after assuming an electric wall across the line of symmetry as shown in Figure 2(b).

Using the conformal mapping technique as explained above, it is possible to show that the even-mode capacitor per unit length ($C_e$) and the odd-mode capacitor per unit length ($C_o$) are

$$C_e = 2c_i \left[ k(k_1)/K(k_1) + c_i K(k_2)/K'(k_2) \right] \quad (3)$$
$$C_o = 2c_i \left[ k(k_1)/K(k_1) + c_i K(k_3)/K'(k_3) \right] \quad (4)$$
$$k_1 = w/(w+2s) \quad (5)$$
$$k_2 = \sinh (\pi w/(2h))/\sinh (\pi (w+2s)/(2h)) \quad (6)$$
$$k_3 = \tanh (\pi w/(2h))/\tanh (\pi (w+2s)/(2h)) \quad (7)$$

$k(k)$ and $K(k')$ are the first kind elliptical integral and its complementary, $c_i$ and $h$ are the dielectric constant of the substrate, and its thickness, respectively, $w$ is the width of the coupled CPW, and $s$ is the slot width. The different dimensions of the structure are depicted in Figures 1 and 2. The mode impedance of each of the two coupled CPW lines at the top and bottom side of the substrate at any of the two modes can be found from Ref. 5

$$Z_{oo} = 1/(c_o \sqrt{C_i C_o}) \quad (8)$$

The subscript $i$ refers to the mode ($e$ for even-mode and $o$ for odd-mode), and $c_o$ is the velocity of light in free space, $C_i$ is the ith mode capacitor, whereas $C_o$ is the ith mode capacitor when the substrate is replaced with free space.

For a certain substrate ($c_i$ and $h$), and required even- and odd-mode impedances ($Z_{ee}$ and $Z_{oo}$), the dimensions of the utilized coupled structure ($w$ and $s$) can be found using (3)–(8).

The physical length of the coupled structure ($l$) is equal to quarter of the effective wavelength calculated at the centre frequency of operation, that is, at 6.85 GHz.

3. RESULTS AND DISCUSSION

To validate the proposed structure and its design procedure, a 45° phase shifter is designed, manufactured using the substrate Rogers RO4003C ($c_i = 3.55, h = 0.508$ mm), and tested. By using the presented design method and with the help of the optimization capability of the software CST Microwave Studio, the dimensions of the structure are found to be: $w = 3.04$ mm, $s = 0.53$ mm, and $l = 5.68$ mm. Concerning the input/ output CPW ports, the width of the central conductor $w_c = 1.1$ mm and the slot between the central conductor and the ground plane $s_i = 0.2$ mm. The overall dimension of the developed device (Fig. 3) excluding the reference transmission line is 2 cm $\times$ 3 cm.

The simulated and measured differential phase shift of the designed 45° phase shifter is shown in Figure 4. The differential phase shift is equal to 45° $\pm$ 2° in the simulation and 45° $\pm$ 3° in the measured results across the band from 3 to 12 GHz. The return loss for the developed phase shifter is better than 30 dB across most of the investigated band. It is better than 10 dB across the band from 3.1 to 11.8 GHz. The insertion loss of the device is less than 0.6 dB according to the simulations and less
than 0.8 dB in the measured results across the band from 3.1 to 10.6 GHz. There is a good agreement between the measured and simulated results.

4. CONCLUSIONS
A compact differential phase shifter for UWB systems has been presented. The proposed filter is based on a single-section broadband-side-coupled CPW structure that can be implemented on a two-sided PCB. A complete design method for the device has been presented and validated.

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RAMAN AMPLIFIED ACCESS NETWORKS WITH PUMP SIGNAL RECYCLING FOR ELECTRICAL POWER CONVERSION

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ABSTRACT: Nowadays, the structure of the passive optical networks is evolving, to keep the network survivability, the powering of some elements in this kind of networks is a key issue. In this work we propose the recycling of the Raman pump remaining optical power at the receiver to be converted to electrical power. With a 24 dBm optical pump signal a 8 dB on/off gain over 25 km of single mode optical fiber was achieved. Also, the recycling of the pump signal generated 18.1 mW of electrical power at the receiver. © 2011 Wiley Periodicals, Inc. Microwave Opt Technol Lett 54:116–119, 2012; View this article online at wileyonlinelibrary.com. DOI 10.1002/mop.26506

Key words: Raman amplifications; power over fiber; passive optical networks

1. INTRODUCTION
The popularity of Internet and the World Wide Web has been leading to the consequent increase of the overall data traffic volume, imposing a plethora of novel questions. This increase on the data traffic increase on the existing telecommunications networks is driving the introduction of the fiber optics into the access layer, using passive optical networks (PON) [1]. PON are becoming a consistent alternative to offer access solutions in a fiber-to-the-home environment [1].

A typical PON is a point-to-multipoint fiber to the user network architecture in which unpowered optical (power or waveform) splitters are used to enable the service to multiple premises. The PON configuration reduces the amount of required fiber and central office equipment when compared with point to point architectures.

Although PON are passive, the electrical power of some network elements, such as the remote node (RN), could enable the protection or restoration of the network in case of failure [2]. However, due to geographical limitations, in some situations, is difficult to electrically power the RN premises. In this work, we propose the use of the optical signal transmitted over the fiber to power the RN.

The concept of RNs optical powering was not extensively exploited yet. However, the recent introduction on the market of photo-voltaic converters to optical fibers (photonic power converters – PPC), with efficiencies up to 15% by JDSU, leads to the possibility of efficiently supply energy to unpowered RNs with optical signals transported in the transmission fiber [3].

Several research groups have been working in the increase of the PONs network range, which could be achieved by the utilization of amplification techniques resilient to power transients, like Raman amplification [4]. The idea of using Raman amplification to provide extended reach and enhanced bandwidth on NGAs networks has been leading to intensive research work. Recently, a passive GPON compatible, with reach extender, using distributed Raman amplifiers over 60 km of fiber has been presented for urban and rural context [5, 6]. Other studies, such as have also demonstrated the feasibility of bidirectional access links using distributed Raman amplification in PON at 10 Gb/s over 80 km [7].

Therefore we propose in this article the possibility to recycle the Raman pump power at the RN to electrical power the node. In Section 2 we describe the theoretical aspects and characterization of the PPC. Section 3 is devoted to the description of the implemented PON scenario with energy recycling, and the experimental results are discussed in Section 4. Finally, in Section 5 the work conclusions are despite.

2. PHOTONIC POWER CONVERTER
A PPC is a photovoltaic cell, allowing the conversion of optical into electrical energy. The devices used to convert optical energy transmitted through an optical fibre, are based on semiconductor materials, such as AlGaAs, GaAs, InGaAs, or InP, with absorption edge bands in the near infrared spectral region, where the optical fiber communications systems operate [8, 9].

The typical diameter of the optical fiber core varies between 8 and 65 μm. These values are extremely small when compared to the active surface of a conventional photovoltaic cell, so, the photo converters require a much larger photocurrent density, relatively to a typical photovoltaic cell. Thus, to effectively convert the light radiation coming from a laser into electricity, high-current photovoltaic cells, based on semiconductor heterostructures, with typically dimension of 1 mm × 1 mm are used. To build sufficient voltage at the PPC terminals, the photovoltaic cell is divided in several segments, electrically isolated and connected in series, to add the voltage for each segment. For a typical cell with six segments of GaAs, it is possible to get an output voltage of 6 V [2]. The grid fingers, shadowing from the electric connection between cells, in this converters, is typically 10%, and may be reduced using a prismatic cover on the cell surface.
Double Microstrip-Slot Transitions for Broadband ±90° Microstrip Phase Shifters

Y. Wang, M. E. Bialkowski, and A. M. Abbosh

Abstract—The letter describes double microstrip-slot transitions for use in planar ±90° phase shifters. The described devices exhibit broadband performance and offer compatibility with ordinary microstrip circuits. Full-wave EM simulation results show a phase shift of ±90° ±7° over the frequency band of 3.1–12.0 GHz when compared with a suitably chosen section of microstrip line. The observed differential phase shift is accompanied by return losses of not less than 14 dB and insertion losses between 0.7 to 1.5 dB in the band 3.1–11.0 GHz. The simulated performance is confirmed by experimental results of ±90°±8° phase shift, return loss not less than 14 dB and insertion loss between 0.5 and 1.8 dB in the frequency band of 3.1–11.0 GHz.

Index Terms—Microstrip circuits, microstrip-slot transition, ultra-wideband (UWB) phase shifter.

I. INTRODUCTION

MICROWAVE digital phase shifters are very useful components in communication sub-systems and radar. Prominent examples include phased array antennas and phase modulators [1]. Modern phase shifters are usually operated using electronic means. They are viewed as two-port networks whose two operational states are controlled by electronic switching elements such as PIN diodes, transistors or MEMS. Their design often commences with fixed configurations which include two circuits, one representing a phase shifting line and the other a reference line.

Many digital phase shifter designs available in the literature are narrowband. The development of wideband systems in the last few decades calls for designs offering a constant differential phase over a broad frequency band [2]–[8]. One prominent example of a broadband phase shifter is the Schiffman phase shifter [2], which uses coupled lines for wideband phasing.

This letter reports the design of an alternative structure in the form of a double microstrip-slot transition that can be used for construction of broadband ±90° phase shifters. The design focuses on the 3.1–10.6 GHz frequency band allocated by US-FCC in 2002 for ultra-wideband applications [9]. It exploits a principle similar to that described in [4], [7], [8], [10], where it was shown that an ultra-wideband phase shift could be obtained using microstrip-slot transitions in a double-layer substrate. In this letter, it is demonstrated that a broadband differential phase shift can be obtained using a double microstrip-slot transition in a single layer substrate. The advantage of this new design is that unlike the one described in [4], it provides compatibility with ordinary microstrip circuits. Also, it offers both +90° and −90° phase shift which cannot be realized using the structure proposed in [4].

II. DESIGN

Two configurations of double microstrip-slot transitions to design wideband ±90° phase shifters in single layer dielectric substrate are presented in Fig. 1.

In these transitions, named in-phase and inverted-phase, microstrip and slotline sections are terminated in virtual open and short circuits. They can be realized using circular, radial or rectangular/linear line or multi-arm lines [10], [13], [14]. The ones shown in Fig. 1(a) use circular shaped stubs. In the two transitions, the signal is launched at one of the microstrip ports. Then it is coupled to a slotline in the ground plane. Next, it travels along the slotline until it is coupled to the second microstrip port. The difference between the in-phase and inverted-phase transitions shown in Fig. 1(a) and (b) is with respect to the direction of the second microstrip line. This direction change introduces a phase shift of 180°. As a result, if the first transition provides the phase shift of $\Delta\phi$, the other transition gives the phase shift of $\Delta\phi = -180°$. This is an advantage of this configuration compared with the design in [4], which does not offer this flexibility.

There are only five design parameters of the in-phase and inverted-phase transitions: radii of two types of stubs, distance between the stubs, microstrip line width and slot width. They are selected using the following simple guidelines.
Once the substrate parameters are specified, radii of the circular microstrip and slotline stubs are chosen to achieve the lowest frequency ($f_{r}$) of operation. The choice of $f_{r}$ introduces the limitation on the upper frequency of operation, $f_{rUP}$, which is the frequency at which the stubs turn into radiating elements [10]. $f_{rUP}$ is approximately equal to $4 \times f_{r}$. Assuming that at $f_{r}$ the open circuit radius is a quarter of the effective slotline wavelength, an estimate for $Rs$ is $R_{s,est} = l/\sqrt{(f/f_{r,GHz} \sqrt{\varepsilon_{eff}})}$ where $\varepsilon_{eff} \approx (1 + \varepsilon_{r})/2$. The short-circuit stub radius, $Rm$, can be estimated in a similar manner and can be determined as $R_{m,est} = l/\sqrt{(f/f_{r,GHz} \sqrt{\varepsilon_{r}})}$ where $\varepsilon_{r}$ approximates $\varepsilon_{eff}$.

For the microstripline having a characteristic impedance of 50 $\Omega$, the slotline width is selected to obtain a characteristic impedance of approximately 84 $\Omega$ [10]. Their widths can be obtained using standard design formulas [11]. There is flexibility in choosing the slotline length $L_{s}$. Here small spacing $L_{s} = \lambda/\varepsilon$ (at 6.5 GHz) is chosen to minimize radiation and conduction losses.

These guidelines are independent of the choice of dielectric constant or thickness of substrate. A larger permittivity is preferred because increasing the slotline width makes the design less prone to manufacturing errors [11]. Also this choice makes the design compact. This affirmative comment does not apply to the phase shifters described in [4]. They are formed by removing two arms of an elliptically shaped slotcoupled patch coupler. The performance of these couplers, and thus the phase shifters, is substrate dependent [12]. Therefore, besides offering $\Delta \phi$ and $\Delta \phi = +\phi_{F}$ phase shift, the substrate independent design is another advantage of the proposed phase shifting structure.

### III. Results

Following the above considerations, the design of transitions aimed at obtaining a constant differential phase shift of $\pm\phi_{F}$ or $\mp\phi_{F}$ in the frequency band of 3.1 to 10.6 GHz when compared with a suitably chosen section of microstripline is undertaken. The design and layout production is aided with Ansoft High Frequency Structure Simulator (HFSS). Rogers RT6010 with thickness 0.635 mm, dielectric constant 10.2 and tangent loss 0.0023 is assumed as a substrate. Fig. 2 shows layouts of the in-phase and inverted-phase microstrip-slot transitions for use in the wideband $\pm\phi_{F}$ phase shifters.

Table I includes parameters of the transitions and the reference microstripline which are used to accomplish the $\pm\phi_{F}$ differential phase shift.

---

**TABLE I**

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<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
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<tr>
<td>Lm</td>
<td>length stub</td>
<td>7 mm</td>
</tr>
<tr>
<td>Lm</td>
<td>length stub</td>
<td>7 mm</td>
</tr>
<tr>
<td>Wm</td>
<td>width stub</td>
<td>3.58 mm</td>
</tr>
<tr>
<td>d</td>
<td>length stub</td>
<td>0.8 mm</td>
</tr>
<tr>
<td>Dm</td>
<td>length stub</td>
<td>5 mm</td>
</tr>
<tr>
<td>Rm</td>
<td>radius stub</td>
<td>1.10 mm</td>
</tr>
<tr>
<td>Rs</td>
<td>radius stub</td>
<td>2 mm</td>
</tr>
<tr>
<td>d</td>
<td>radius stub</td>
<td>0.5 mm</td>
</tr>
<tr>
<td>tref</td>
<td>length stub</td>
<td>34 mm</td>
</tr>
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<td>34 mm</td>
</tr>
<tr>
<td>tref</td>
<td>radius stub</td>
<td>0.70 mm</td>
</tr>
</tbody>
</table>

Fig. 2. Layouts of (a) in-phase and (b) inverted-phase double microstrip-slot transitions.

Fig. 3. Photograph of the manufactured reference line and double in-phase and inverted-phase microstrip-slot transitions, top (a) and bottom (b) view.

Fig. 4. Simulated and measured magnitude characteristics of the designed in-phase and inverted phase transitions for use in $\pm\phi_{F}$ phase shifter.
Following the design stage, the two varieties of the phase shifter are manufactured and experimentally tested. Fig. 3 shows a photograph of the manufactured devices.

Fig. 4 presents the simulated and experimental amplitude characteristics of the two transitions for construction of the two phase shifters. For both the in-phase and inverted-phase transition, the simulated return loss is greater than 14 dB and insertion loss increases from 0.7 to 1.5 dB in the band of 3.1–11 GHz. The measured return loss is greater than 17 dB for the in-phase and not less than 14 dB for the inverted-phase, while the insertion loss is between 0.5 and 1.8 dB for the two transitions in the frequency band of 3.1–11 GHz. The observed increased insertion loss is caused by residual radiation. This type of loss is also observed in designs in [2] and [4]. It can be minimized by introducing a conducting enclosure [2].

Fig. 5 shows the simulated and measured results of the phase characteristics for the two phase shifters. The experimental phase shift is $\phi \pm 90^\circ$ over the band of 3.1–12 GHz and is comparable with the simulated one of $\phi \pm 7^\circ$.

IV. CONCLUSION

In this letter, the design of broadband double microstrip-slot transition for use in $+\phi$ and $-\phi$ phase shifters has been presented. The phase shifters developed using these transitions are fully planar and compatible with microstrip circuits. They feature good amplitude and phase shift characteristics exceeding the initially specified 3.1–10.6 GHz band. They are attractive from the point of view of ease of fabrication without limitations of substrate choice.

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Tunable phase shifter employing variable odd-mode impedance of short-section parallel-coupled microstrip lines

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Abstract: A complete design method for a tunable phase shifter that employs a short section of parallel-coupled microstrip lines is presented. In the proposed method, the variation in the phase is achieved by changing the odd-mode impedance of parallel-coupled microstrip lines using a varactor diode that is connected between them. A derived theoretical model shows that a unit-cell phase shifter of around one-tenth of the guided wavelength can be utilised to achieve a continuously tunable phase range in excess of 90° depending on the required bandwidth and acceptable insertion loss. The proposed method shows that in order to achieve the required tunable phase range across a wideband, high even-mode impedance is needed. A slotted ground structure is utilised underneath the coupled structure to realise that target. The proposed method is validated by building a phase shifter that has a very small length of around one-twentieth of the guided wavelength with 45° tunable phase range and about 1 dB insertion loss across the band from 2 to 2.5 GHz.

1 Introduction

Tunable phase shifters are used to control the phase characteristics of signals to be processed in any system. Therefore phase shifters are key devices in many microwave systems, such as phased antenna array, mobile satellite systems, microwave instrumentation, modulators, noise cancellation systems, frequency converters and wireless local area networks employing multiple-input multiple-output technology [1–21].

According to the industry outlook, there are three main challenges facing the wide use of tunable phase shifters. The first one comes from the cost. It is estimated that the cost of tunable phase shifters represents nearly half of the cost of the entire electronically scanned array used in communication systems [2]. The other important challenge, especially in modern communication systems with large dynamic ranges, is the level of insertion loss caused by the utilised phase shifters. It is a crucial parameter that defines the overall performance of those systems. The insertion loss of phase shifters used in a transmitter causes a significant reduction in the level of transmitted power, whereas it causes degradation in the signal-to-noise ratio when the phase shifter is part of a receiver [3]. Both these effects reduce the dynamic range significantly of even the best designed systems. The third challenge is the capability to achieve the same phase change at a certain applied control signal across the whole band of interest. This factor is especially important when using the tunable phase shifters in wideband systems.

In general, phase shifters can be classified as fixed [4–8], digital [9–12] or analogue phase shifters [13–21]. The fixed phase shifter uses loaded or coupled lines to change the phase by a constant value across a certain band. The digital type can change the phase in a certain limited number of values with the help of different types of switching elements. The third one, which is called analogue phase shifters and is the topic of this paper, can change the phase continuously depending on the control signal. Therefore analogue phase shifters offer unlimited resolution, and thus, they have the widest range of applications.

The most common type of analogue phase shifters that uses coupled structures is the reflection type [13–18]. It depends on directional couplers to split the input signal into multiple orthogonal signals that are combined at the output. The variable phase shift is realised by changing the amplitude of those orthogonal signals before their combination. This operation requires a perfectly even power split, 90° phase difference across the required band and parasitics-free diodes. Since those factors cannot be maintained in a perfect manner across a wideband, the result is usually a high insertion loss.

Some of the analogue phase shifters are built using different types of thin-film technologies [15–18]. Apart from the complexity and cost of those technologies, they have the clear benefit of achieving a compact size and easy of integration with other devices. Since the thin-film-based phase shifters generally suffer from more than 5 dB insertion losses, some of the proposed structures include embedded amplifiers to compensate for those losses [15, 18]. However, the overall noise figure of those devices is more than 10 dB indicating a limited dynamic range and degraded noise immunity. It is well known that the noise
added owing to the use of phase shifters with high insertion loss cannot be removed by just increasing the gain of the following stages [22].

In this paper, a short section of parallel-coupled microstrip lines is utilised to develop a continuously tunable phase shifter. The phase variation is achieved by changing the odd-mode impedance of the coupled structure. A varactor diode is connected for that purpose between the two coupled lines. A complete design method is derived to show the relation between the length of the coupled structure, the even-mode impedance, range of odd-mode impedance, the fractional bandwidth and the tunable phase range. The proposed approach is implemented by building a 0.06A length phase shifter that has a continuous tunable phase of 45° with less than 1.2 dB of insertion loss across the band from 2 to 2.5 GHz.

2 Theory

Assume that a coupled structure of arbitrary length \( l \) is connected in the manner shown in Fig. 1. A varactor diode is connected between the two coupled lines. The line of symmetry for the coupled structure depicted inside the dotted square behaves as an E-wall in the odd mode and H-wall in the even mode. Thus, it is easy to conclude that the presence of this varactor has no effect on the even-mode impedance, whereas it has a certain impact on the odd-mode impedance \( Z_{oo} \), depending on the value of the varactor’s capacitor \( C \).

Using the signal flow diagrams of four-port devices [23], it is possible to show that for arbitrary \( l \), \( C \), \( Z_{oe} \) and \( Z_{oo} \), the outgoing signals \( (b_i) \) can be calculated for the structure of Fig. 1 as follows

\[
b = [I - SF]^{-1}Sa
\]  

(1)

\( b \) is the vector representing the signal out of each port, \( I \) the identity matrix and \( a \) the vector of input signals from outside sources. For the structure of Fig. 1, values of \( a_i \), where \( i \) refers to the port number, are

\[
a_1 = 1, \quad a_2 = a_3 = a_4 = 0
\]  

(2)

\( \Gamma \) is the diagonal matrix representing the reflection coefficients at the four ports. All the elements of the matrix are zero except for the diagonal elements where the values derived from the structure of Fig. 1 are

\[
\Gamma_1 = \Gamma_4 = 0, \quad \Gamma_2 = \Gamma_3 = 1
\]  

(3)

\( S \) is the scattering matrix with elements are calculated using

\[
S_{11} = (S_{11e} + S_{11o})/2
\]  

(4)

\[
S_{21} = (S_{21e} + S_{21o})/2
\]  

(5)

\[
S_{31} = (S_{31e} - S_{11o})/2
\]  

(6)

\[
S_{41} = (S_{21e} - S_{21o})/2
\]  

(7)

\[
S_{11e} = \frac{j(Z_{oe}/Z_o - Z_{oo}/Z_{oo}) \sin \beta l}{2 \cos \beta l + j(Z_{oo}/Z_o + Z_{oe}/Z_{oo}) \sin \beta l}
\]  

(8)

\[
S_{11o} = \frac{j(Z_{oe}/Z_o - Z_{oo}/Z_{oo}) \sin \beta l}{2 \cos \beta l + j(Z_{oo}/Z_o + Z_{oe}/Z_{oo}) \sin \beta l}
\]  

(9)

\[
S_{21e} = \frac{2}{2 \cos \beta l + j(Z_{oo}/Z_o + Z_{oe}/Z_{oo}) \sin \beta l}
\]  

(10)

\[
S_{21o} = \frac{2}{2 \cos \beta l + j(Z_{oo}/Z_o + Z_{oe}/Z_{oo}) \sin \beta l}
\]  

(11)

\( \beta \) is the phase constant in the medium of the coupled structure.

The important parameters that define the performance of the structure shown in Fig. 1 are the phase of the output signal \( \Phi \), the reflection coefficient at the input and output ports \( (S_{aa} \) and \( S_{bb} \)) and the transmission coefficient from the input to the output port \( (S_{ba}) \). Those parameters can be calculated for the structure of Fig. 1 after solving (1) as follows

\[
S_{aa} = S_{bb} = b_1/a_1, \quad S_{ba} = b_2/a_1, \quad \Phi = \text{angle}(S_{ba})
\]  

(12)

The performance of the proposed structure concerning the insertion loss and the phase of the output signal are calculated using (1)–(12) for a wide range of values for the mode impedances. Snapshots from the calculations are shown in Fig. 2 for even-mode impedances from 100 to 200 \( \Omega \) and for insertion losses that are equal to or less than 1 and 2 dB. The procedure used in the calculations is to use (1)–(12) to find the phase shift for the assumed \( Z_{oo} \) and all

![Fig. 1 Coupled lines with varactor diode that changes the odd-mode impedance](image1)

![Fig. 2 Variation of the maximum tunable phase range with length of the coupled structure](image2)
the possible values of $Z_{oo}$ that keep the insertion loss calculated from $S_{21A}$ equal or below the target.

From Fig. 2, it is possible to conclude that the maximum tunable phase range can be obtained when $l \approx n\lambda/4 \pm 0.125\lambda$ ($n = 0, 1, 2$ etc. and $\lambda$ is the guided wavelength). For a compact structure, the maximum tunable phase range ($\Delta \Phi$), which increases with $Z_{oo}$, can practically be obtained at $l \approx 0.125\lambda$. Also, Fig. 2 reveals that the coupled structure cannot be used as a tunable phase shifter at $l = n\lambda/4$ by simply varying the values of any of the mode impedances as there is no change in the phase at those lengths.

Besides $\Delta \Phi$, the other parameter, which is critically important when designing the proposed tunable phase shifter, is the range of values for the $Z_{oo}$ needed to achieve a certain value for $\Delta \Phi$. This parameter is limited by the feasible range for the varactor’s junction capacitor $C_v$. In general, it is required to have a low range of variation in the odd-mode impedance as this means a low variation in $C_v$ and consequently a low variation in the applied biasing voltage, or a low-cost diode. For that reason, the normalised range of the odd-mode impedance ($R_n$), which is the ratio of the maximum to the minimum odd-mode impedance normalised by $\Delta \Phi$ ($R_n = (Z_{oomax}/Z_{omin})/\Delta \Phi$) is calculated for the results of Fig. 2 and presented in Fig. 3.

It is interesting to see that the minimum value for $R_n$, which means a minimum range of variation in $Z_{oo}$ per degree of phase change, occurs at $l \approx 0.05\lambda$. A close inspection of the minimum values for $R_n$ in Fig. 3 shows that they decrease slightly with increasing $Z_{oo}$. Apart from that, it is possible to conclude from Fig. 3 that $R_n$ generally has very small dependence on $Z_{oo}$. By comparing the results of Fig. 2 with those of Fig. 3, it is possible to say that if the requirement is a low range of $\Delta \Phi$, the most compact design is obtained with $l \approx 0.05\lambda$. However, if $\Delta \Phi$ is required to be as large as possible, the design in this case requires $l \approx 0.125\lambda$.

The other important factor in the design is the fractional bandwidth that can be achieved with a certain acceptable deviation in the tunable phase across that band. The theoretical model (1)–(12) is used to calculate the maximum value of $\Delta \Phi$ as a function of the required fractional bandwidth assuming that the maximum acceptable deviation in $\Delta \Phi$ is 10% and that the maximum insertion loss across the band of interest is 1, 2 and 3 dB. The results are shown in Fig. 4. As a practical precaution adopted in the calculations, the odd-mode impedance was allowed to take values that are larger than 4 $\Omega$. Extremely small values for $Z_{oo}$ require a very narrow gap between the coupled lines and/or a special and expensive varactor diode that has a wide range for $C_v$. It is to be noted that an extremely narrow gap causes a difficulty in the manufacturing process and in connecting the varactor diode between the coupled lines.

From Fig. 4, it can be concluded that there is an inverse relationship between the maximum tunable phase range and the required fractional bandwidth. Moreover, the achievable $\Delta \Phi$ increases with the value of the acceptable insertion loss. Since the length of the coupled structure needed to achieve the results of Fig. 4 is quite small (0.05$\lambda$ to 0.1$\lambda$), wide bandwidths with large $\Delta \Phi$ can be obtained by cascading several sections as required.

As an example for the design procedure, assume designing a tunable phase shifter with $\Delta \Phi = 45^\circ$ across the band from 2 to 2.5 GHz, that is 22.2% fractional bandwidth. For practical reasons, the minimum value for the odd-mode impedance is set at 4 $\Omega$. Two designs are required; one for an insertion loss that is less than 1 dB, whereas the other is for less than 2 dB insertion loss. As a first step in the design, Fig. 4 is inspected to make sure that a single section can be used to meet the design requirements. Using the derived model (1)–(12) in an iterative MATLAB code aimed at finding the optimum values for the even-mode impedance, required range of the odd-mode impedance and length of the coupled structure, the following values are obtained. For the 2 dB insertion loss design, $l = 0.05\lambda$, $Z_{oo} = 350$ $\Omega$, $Z_{oo} = 4 – 18$ $\Omega$. For the 1 dB insertion loss design, $l = 0.06\lambda$, $Z_{oo} = 320$ $\Omega$, $Z_{oo} = 4 – 24$ $\Omega$. The design can also be achieved with almost the same performance using lower values for the even-mode impedance. However, the design in that case requires a larger range for the required odd-mode impedance, which means the need for a special varactor diode, and a larger length for the coupled structure, which violates the target of building a compact device.

As can be seen from the calculated values of the design parameters, the strict requirement of as low as 1 dB insertion loss
loss across the whole band requires a certain price to be paid. That price includes a larger size and higher range of odd-mode impedance, which could mean a more expensive varactor or/and larger range of applied biasing voltage. The phase performance of the two designs relative to the minimum $Z_{\text{oe}} = (4 \Omega)$ case of each design is shown in Figs. 5 and 6. The tunable phase range for the first design with 2 dB insertion loss is $\Delta \Phi = 45 \pm 4.5^\circ$, whereas it is $45 \pm 5.5^\circ$ for the 1 dB insertion loss design. Thus, relaxing the insertion loss requirements helps reducing the deviation in the phase shift across the required band.

3 Design

As concluded from the calculated results in Fig. 2, the proposed tunable phase shifter is required to have high even-mode impedance and a range of low values for the odd-mode impedance. If a parallel-coupled microstrip structure is utilised to build the phase shifter, the high value for $Z_{\text{oe}}$ can be achieved by using a slotted ground plane, which results in an increase in the even-mode impedance without significant impact on the odd-mode circuit [25]. Concerning the requirement of a range of low odd-impedance values, it can be achieved via the selection of a suitable varactor diode with relatively high junction capacitor. The implemented structure of the tunable phase shifter is shown in Fig. 7. It includes the biasing circuit for the varactor diode. There is a radio frequency choke (RFC) of 10 $\mu$H to isolate the microwave signal from the biasing line, and a 1 nF chip capacitor to block the biasing voltage from the input and output ports.

In order to find the initial dimensions of the coupled structure and the required range of the needed junction capacitor for the varactor diode, a quasi-transverse electromagnetic propagation is assumed for the structure of Fig. 7. Thus, the even- and odd-mode impedances of the coupled lines are determined from the effective capacitances per unit length of the lines and the phase velocity in the utilised medium [24]. With the help of the conformal mapping technique [25], the following results that show the relation between the mode impedances and the dimensions of the coupled structure can be obtained

$$Z_{\text{oe}} = \frac{60\pi K(k_1)}{\sqrt{\varepsilon_r K'(k_1)}}$$

$$Z_{\text{oo}} = \frac{60\pi\sqrt{(\varepsilon_r + 1)/2}}{v_n (k_1/k_2) + K'(k_2)/K(k_2) + K(k_2)/K'(k_2) + C_{\text{var}}/\varepsilon_r}$$

$$k_1 = \frac{1 + \exp(-\pi w_m/(2h))}{1 + \exp(-\pi w_m/(2h))}$$

$$k_2 = \frac{\tanh(\pi s/(4h))}{\tanh(\pi s/(4h))}$$

$$k_3 = \frac{s}{s + 2w_m}$$

where $K$ and $K'$ are the first kind elliptical integral and its complementary, respectively, of the parameters $k_1$, $w_m$ the width of any of the coupled lines, $w_s$ the width of the slot in the ground plane and $s$ the width of the gap between the coupled lines. The parameters $w_m$, $w_s$, and $s$ are shown in Fig. 7. $C_{\text{var}} = C_s/l$, junction capacitor of the varactor ($C_s$) per unit length of the coupled structure and $s_t$ and $h$ are the dielectric constant and thickness of the substrate, respectively. The use of slot in the ground underneath the coupled structure leads to a significant increase in the even-mode

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**Fig. 5** Variation of the phase with frequency at different values of the odd-mode impedance with insertion loss $\leq 2$ dB

**Fig. 6** Variation of the phase with frequency at different values of the odd-mode impedance with insertion loss $\leq 1$ dB

**Fig. 7** Schematic diagram showing the practical implementation of the proposed device
impedance. This conclusion can be verified by the presence of the parameter $w_s$ in (15) and thus (13). The removal of the ground plane underneath the coupled structure has a small effect on the odd-mode impedance because $k_1$ has negligible effect on (14) compared with $k_2$ and $k_3$. On the other hand, as indicated by (13) and (14), $C_vn$ has a direct effect on the odd-mode impedance of the coupled lines without any impact on the even-mode impedance.

4 Results and discussions

A unit-cell tunable phase shifter that can be tuned by $45^\circ$ with less than 1 dB insertion loss across the band from 2 to 2.5 GHz, that is, fractional bandwidth $= 22.2\%$, was designed and fabricated. The band 2–2.5 GHz is chosen for the design as it is increasingly used in the modern technique to build high-capacity wireless local area networks employing multiple-input multiple-output front-ends [26]. The key component in that type of front-ends is the phased array that is controlled by tunable phase shifters. Rogers RT6010 ($\varepsilon_r = 10.2$, $h = 0.635$ mm) was used as the substrate.

As explained earlier, the theoretical model (1)–(12) is used to find the required value for $Z_{oo}$ (320 $\Omega$), range of values for $Z_{oo}$ (4–24 $\Omega$) and length of the coupled structure $l$ (0.06 $\lambda$). The substitution of those values in (13)–(17) results in the following dimensions for the coupled structure and range of varactor’s capacitor: $w_m = 0.29$ mm, $s = 0.25$ mm, $l = 3$ mm, $w_s = 7$ mm and $C_v$ from 0.45 to 4.3 pF.

To verify the reliability of the proposed design approach, the performance using those calculated values is simulated using the software tool CST Microwave Studio. The results concerning the phase performance relative to the case of zero diode biasing (maximum $C_v$ and thus minimum $Z_{oo}$) and the amplitude of the S-parameters are shown in Figs. 8 and 9. The achieved tunable phase is equal to $47 \pm 5^\circ$ in the band from 2 to 2.5 GHz. The insertion loss across that
band is less than 1.75 dB. These results prove that the presented design approach gives reasonable estimation for the values of the design parameters. However, the performance of the phase shifter still needs an improvement. Thus, the designed parameters were optimised using the software tool and the final values are: \( w_m = 0.3 \text{ mm}, \quad s = 0.21, \quad l = 2.9 \text{ mm}, \quad w_s = 5.8 \text{ mm}, \quad C_v \) from 0.6 to 6 pF. The developed device with the optimised dimensions is shown in Fig. 10. To achieve the required range for \( C_v \), a GaAs hyperabrupt varactor diode with a maximum biasing voltage of 12 V is used.

The performance of the developed device was verified via simulations and measurements. The simulated and measured phase at different biasing voltages relative to the case of zero biasing (maximum \( C_v \)) is shown in Fig. 11. It is clear that using a short coupled section of 2.9 mm with a single varactor diode enables controlling the phase by a range of 45°. This value for the range of phase change is maintained across the whole investigated band. The maximum fluctuation in the achieved phase is \( \pm 2^\circ \) in the simulated and \( \pm 4.5^\circ \) in the measured results across the whole band. If the simulated and measured results are compared with those predicted using the derived theoretical model (Fig. 6), it can be concluded that the overall phase performances of the three results are in good agreement.

The amplitudes of the S-parameters for the developed device are shown in Fig. 12. The simulated and measured insertion loss is less than 1.2 and 1.4 dB, respectively, across the band from 2 to 2.5 GHz. This result is very close to the required value (1 dB) in the design criteria. The simulated and measured S-parameters with frequency agree well with each other. The small discrepancy between them is due to the tolerance of the milling machine used in the fabrication and the parasitic elements of the utilised varactor diode.

5 Conclusion

A complete design method for a tunable phase shifter has been presented. The device utilises a variable odd-mode impedance of a short section of parallel-coupled microstrip lines to achieve the required performance. The variation in the odd-mode impedance is implemented using a varactor diode that is connected between the two coupled lines. The proposed method can be used to build a unit-cell phase shifter that has a length of around one-tenth of the guided wavelength for a tunable phase range in excess of 90°. The proposed method is validated by building a 0.06\( \lambda \) length phase shifter that covers the band from 2 to 2.5 GHz with a tunable phase range of 45° and less than 1.2 dB insertion loss.

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Compact Tunable Reflection Phase Shifters Using Short Section of Coupled Lines

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Abstract—In the design of reflection-type phase shifters, the coupler that represents the shifter’s backbone is usually assumed to be a quarter-wavelength 3-dB coupler. In this paper, a derived theoretical model shows that, for certain values of the odd- and even-mode impedances, a coupled structure with a length that is less than one tenth of a wavelength is sufficient to build a high-performance reflection phase shifter. The presented analysis indicates that reflection phase shifters can be designed with a more compact size and larger phase range compared with the conventional method of using a quarter-wavelength 3-dB coupler. However, the required odd-mode impedance in the proposed design is low ($\approx 10 \Omega$), whereas the required even-mode impedance is high ($\approx 200 \Omega$). To realize those impedances when using parallel-coupled lines, slotted ground and shunt chip capacitor are used. The proposed design is supported by full-wave electromagnetic simulations and measurements. The simulated results show that a quarter-wavelength coupled structure achieves 255° phase range across 36% fractional bandwidth with less than 1-dB insertion loss and more than 10-dB return loss. In another design, a full-cycle phase range is obtained with less than 1.5-dB insertion loss across the same band by using two $\lambda/4$ coupled sections. A manufactured prototype for a full-cycle phase range validates the simulation results and, thus, the proposed method.

Index Terms—Analog phase shifter, reflective phase shifter, tunable phase shifter.

I. INTRODUCTION

UNABLE phase shifters are key devices in many microwave systems, such as phased arrays, satellite systems, microwave instrumentations, modulators, noise cancellation systems, frequency converters, and, recently, wireless local area networks (WLANs) employing multiple-input multiple-output (MIMO) technology [1]–[11].

The phase shifters are required to have compact size, low cost, and low insertion loss across the required bandwidth. The size of the phase shifters has become a crucial parameter in their design, especially since the recent adoption of the MIMO technology in the design of mobile handsets due to the limited available space. Moreover, the cost of the utilized phase shifters in the handsets and other portable devices should be as low as possible for obvious economical reasons. In addition, the level of the insertion loss caused by the utilized phase shifters is the key factor that defines the overall performance of modern communication systems with large dynamic ranges. A significantly high insertion loss of phase shifters used in a transmitter causes a significant reduction in the level of transmitted power, whereas it causes a serious degradation in the signal-to-noise ratio (SNR) when the phase shifter is part of a receiver. Both of those effects reduce the dynamic range significantly of even the best designed systems.

The most common type of analog phase shifters is the reflection-type [1]–[15]. It uses quarter-wavelength 3-dB quadrature couplers to split the input signal into two orthogonal signals that are reflected back and combined at the output. The variable phase shift is realized by changing the phase and amplitude of those orthogonal signals before their combination.

Reviewing the literature shows that all of the papers that dealt with the reflection-type phase shifter assume by default that the coupler, which is the backbone of the phase shifter, is quarter-wavelength 3-dB quadrature coupler. In the design of quadrature couplers as a standalone device, it is well understood that a quarter-wavelength coupled structure is needed. Moreover, the odd- and even-mode impedances ($Z_{im}, Z_{en}$) in the couplers are chosen according to the formula $\sqrt{Z_{im} Z_{en}} = Z_0$, where $Z_0$ is the input/output ports impedance. However, the final structure and required performance for the couplers and phase shifters as standalone devices are not exactly the same. The quadrature couplers are needed to have a certain coupling factor with 90° phase shift between its two output signals, whereas the phase shifter that has one output should achieve a specific phase range with a certain low insertion loss across a certain band. Thus, the reasonable questions concerning this matter are, do we really need a coupling length of quarter wavelength to achieve the required performance for the phase shifter? Are the quarter-wavelength and 3-dB coupling the optimum choices for the design of reflection-type phase shifters? Does the aforementioned relation between the mode impedances give the best performance for the phase shifter across a certain band? It could be possible that different criteria from those of the quadrature coupler are needed for the mode impedances and length of the coupled structure when designing a reflection phase shifter.

In this paper, a complete theoretical analysis is presented to show that a coupled structure of length that is less than one tenth of a wavelength is sufficient to build a high-performance reflection-type phase shifter. Moreover, the short-length coupled structure is actually the optimum choice to realize the highest possible phase range for a certain varactor diode or reflection load in general. The derived model is supported by full-wave electromagnetic simulations of two designs. Also, a prototype with full-cycle phase range is built according to the proposed method and successfully tested.
II. THEORY

A diagram showing a reflection-type phase shifter utilizing a coupler of length \( l \) is shown in Fig. 1. To change the phase, a variable load \( (Z_l) \) is connected between the two outputs of the coupled structure and the ground. Instead of using quarter-wavelength coupler, assume that the coupled structure’s length \( l \) is a variable to be found for a maximum phase range across a certain band. Also, the mode impedances \( (Z_{ee}, Z_{oe}) \) of the coupled structure are assumed to be independent variables to be optimized for an acceptable performance (insertion loss and return loss) across the required band. In other words, the mode impedances are not assumed by default to be selected to achieve a 3-dB coupling and to be related with each other and with the input/output ports impedance \( (Z_o) \) via the relation \( \sqrt{Z_{ee}Z_{oe}} = Z_o \).

Using the signal flow diagrams of four-port devices [16], it is possible to show that, for arbitrary \( l, Z_o \) and \( Z_{oo} \), the outgoing signals \( (b_i) \) can be calculated for the structure of Fig. 1 as follows:

\[
b = (I - S\Gamma)^{-1}S\alpha \tag{1}
\]

where \( b \) is the vector representing the signal out of each port; in the following analysis, \( b_i \) refers to the output from port \( i \), \( I \) is the identity matrix, \( \alpha \) is the vector of input signals from outside sources which, for the structure of Fig. 1, are

\[
a_1 = 1 \quad a_2 = a_3 = a_4 = 0 \tag{2}
\]

where \( a_i \) is the input signal to port \( i \). \( \Gamma \) is the diagonal matrix representing the reflection coefficients at the four ports. All of the elements of the matrix are zero except the diagonal elements, which are given for the structure of Fig. 1 by

\[
\Gamma_1 = \Gamma_4 = 0, \quad \Gamma_2 = \Gamma_3 = \frac{Z_4 - Z_{oo}}{Z_1 + Z_{in}} \tag{3}
\]

where \( Z_{in} \) is the input impedance of the coupled structure looking from the load’s side. In (3), the input and output ports \( (1 \) and 4) are assumed to be perfectly matched. For a general coupled structure that has a length \( (l) \), even-mode impedance \( (Z_{ee}) \), odd-mode impedance \( (Z_{oe}) \), medium with phase constant \( (\beta) \), \( Z_{ic} \) is given as [17]

\[
Z_{in} = Z_o + \frac{2}{Z_{ic}^2 + Z_{oc}^2} \frac{Z_{in}Z_{in} - Z_o^2}{Z_{oc} + Z_{ic} + 2Z_o} \tag{4}
\]

\[
Z_{in}^w = Z_{vo}Z_{o+\beta}Z_{at}Z_{at} \tan(\beta l) \frac{Z_{at}}{Z_{o+\beta}Z_{at} \tan(\beta l)} \tag{5}
\]

Scattering matrix with elements are calculated using the odd-even mode approach [17]

\[
S_{11} = \frac{S_{11e} + S_{11e}}{2} \tag{7}
\]

\[
S_{21} = \frac{S_{21e} + S_{21e}}{2} \tag{8}
\]

\[
S_{41} = \frac{S_{41e} + S_{41e}}{2} \tag{9}
\]

\[
S_{41e} = \frac{j(Z_{ee}/Z_o - Z_o/Z_{ee}) \sin(\beta l)}{2 \cos(\beta l) + j[Z_{ee}/Z_o + Z_o/Z_{ee}] \sin(\beta l)} \tag{10}
\]

\[
S_{11o} = \frac{j[Z_{oo}/Z_o - Z_o/Z_{oo}] \sin(\beta l)}{2 \cos(\beta l) + j[Z_{oo}/Z_o + Z_o/Z_{oo}] \sin(\beta l)} \tag{11}
\]

\[
S_{21e} = \frac{2[2 \cos(\beta l) + j[Z_{ee}/Z_o + Z_o/Z_{ee}] \sin(\beta l)]}{2 \cos(\beta l) + j[Z_{ee}/Z_o + Z_o/Z_{ee}] \sin(\beta l)} \tag{12}
\]

Equations (1)–(15) are analyzed using a suitable MATLAB code to find the maximum achievable phase range \( (\Delta\Phi) \) as a function of the length of the coupled structure and the mode impedances. In order to limit the calculations to practical values, the mode impedances are allowed to take any value within the range 4 to 400 \( \Omega \). Concerning the varactor diodes, \( C_{min} \) is assumed to be 0.2 pF, whereas the maximum varactor’s capacitor ratio \( R_c \) is assumed to be 10. In order to include only the useful results, it is assumed that the minimum acceptable return loss is 10 dB.

The variation of the maximum achievable \( \Delta\Phi \) as a function of the length of the coupled structure is shown in Fig. 2 for three values of the fractional bandwidth. The fractional bandwidth is defined here as the band with more than 10-dB return loss. It is clear from the results that a coupled structure with a length of around one tenth of a wavelength (\( \lambda \) is calculated at the center of the band) gives the maximum achievable \( \Delta\Phi \). For low values of the fractional bandwidth, the required length is slightly lower than 0.1\( \lambda \), whereas it is slightly larger than 0.1\( \lambda \) for large fractional bandwidths.

The other important factor indicated by the results in Fig. 2 is that the phase range achieved using a short section of coupled structure is larger than the achievable value when using quarter wavelength coupler. It is possible by using the derived method for a traditional reflection-type phase shifter [18] that the phase range for a 10% fractional bandwidth is 109.8°.
The same value is also obtained by using the presented model (1)–(15) when the mode impedances are limited according to the formula \( \sqrt{Z_{\text{in}}R_{\text{c}}} = Z_{\text{o}} \) as needed in the quarter-wavelength 3-dB coupler. Thus, for 10% fractional bandwidth, the increase in the phase range by using \( l = 0.1\lambda \) with proper mode impedances is 30% compared with the phase range when following the traditional approach of using quarter-wavelength 3-dB coupler. For larger fractional bandwidths, Fig. 2 indicates that the increase in the phase range from using the proposed method becomes smaller, but the size of the needed structure is still more compact than the traditional design of reflection-type phase shifters by using quarter-wavelength 3-dB couplers.

The required even-mode impedances to achieve the phase range for each of the cases depicted in Fig. 2 are shown in Fig. 3. It is clear that, if a short coupled structure is used in the design of a reflection phase shifter, large values for the even-mode impedance are needed. This result does not impose any threat to the success of the proposed technique as several techniques can be employed to realize the required mode impedances, as will be shown later. Concerning the required odd-mode impedances to achieve the results of Fig. 2, the values are depicted in Fig. 4. It is obvious that the maximum \( \Delta \Phi \), which occurs around \( l = 0.1\lambda \), requires reasonable and easily implemented values for \( Z_{\text{in}} \).

If \( \Delta \Phi \) is required to be larger than the values depicted in Fig. 2, larger values for the varactor’s capacitor ratio \( R_{\text{c}} \) are needed. For example using \( R_{\text{c}} = 20 \) enables achieving 180° phase range with \( l = 0.1\lambda \). However, this type of varactor is, at the least, not available to the author. Since the analysis presented in this paper is limited to the practical and feasible values, this option is disregarded. Thus, other options can be used to extend the phase range while keeping the size compact.

One of the possible methods is to connect an inductor of suitable value \( L_{\text{i}} \) in series with the varactor diode [3]. Thus, the load impedance \( Z_{\text{L}} \) depicted in Fig. 1 is equal in this case to \( Z_{\text{L}} = j(\omega L - 1/\omega C_{\text{c}}) \). This value for \( Z_{\text{L}} \) is used in the MATLAB code aimed at solving (1)–(15) to find the maximum achievable phase range. In the calculations, the practical limitations on the mode impedances (more than or equal to 4 \( \Omega \) and less than or equal to 400 \( \Omega \) and varactor’s capacitor ratio \( R_{\text{c}} \leq 10 \)) are imposed. Also, the calculations are limited to the cases with more than 10-dB return loss. The results of the calculations are shown in Fig. 5 for a reasonable range of values for \( l \).

From the presented results, it is clear that, as expected, using an inductor in series with the varactor diode increases the phase range significantly. With the practical imposed limits on \( R_{\text{c}} \), \( C_{\text{min}} \), \( Z_{\text{in}} \), and \( Z_{\text{L}} \), a maximum phase range is achieved at a certain value for \( l \). Above and below that value, the achievable phase range decreases.

The achievable phase range from using the traditional approach for the design, i.e., using a quarter-wavelength 3-dB coupler, is also included in Fig. 5 for the same \( l \) and \( R_{\text{c}} \) range of values. The phase range in this case for a 10% fractional bandwidth is calculated using the presented general model (1)–(15) and [18, eq. 14.16] for a phase shifter that uses a quarter-wavelength 3-dB coupler and a series combination of varactor diode and inductor as a reflective load. It is clear from the results that the traditional design method has significantly lower phase range compared with the proposed compact design with optimized mode impedances. It is also clear that the traditional...
design method [18] is a special case of the proposed general method in this paper, as both of them give exactly the same estimation of the phase range.

As a comparison with using the conventional a quarter-wavelength 3-dB coupler, Fig. 5 shows that it is possible, for example, to use a coupled structure with \( l = 0.055 \lambda \) to achieve more than 300° phase range across a 10% fractional bandwidth. If a traditional quarter-wavelength 3-dB coupler is used with the same varactor, i.e., \( H_c = 1 \), and an optimum inductor, it can only achieve about 240° phase range.

The important parameter that is related to the main target of this work is the optimum length of the coupled structure needed to achieve the maximum phase range of Fig. 5. The variation of the length needed to realize the maximum phase range for each value of the inductor is depicted in Fig. 6 after using the MATLAB code to solve (1)–(15). None of the investigated cases shows that a quarter-wavelength coupler is the optimum choice. To the contrary, the results of Fig. 6 indicate that a short-section coupler is the one that is able to achieve the largest phase range if the mode impedances of the coupler are chosen properly. It is obvious also from Fig. 5 that the required \( l \) decreases with increasing the value of the inductor.

The required even- and odd-mode impedances to achieve the phase range for each of the cases depicted in Fig. 5 are shown in Figs. 7 and 8. To achieve the maximum possible phase range across the required bandwidth, the odd-mode impedance of the short coupled structure needs to be around 10 \( \Omega \), whereas the even-mode impedance needs to be around 200 \( \Omega \). Thus, the optimized short-section design requires higher even-mode impedance and lower odd-mode impedance than the values needed in the traditional design method.

### III. DESIGN

To prove the validity of the presented design approach, two phase shifters are designed to cover the frequency band from 1.8 to 2.6 GHz, i.e., 36% fractional bandwidth with 10-dB return loss as a reference. This band is chosen for the design as it is increasingly used in the modern technique to build high-capacity wireless local area networks employing multiple-input
multiple-output (MIMO) front-ends [19]. The key component in that type of front-ends is the phased array that is controlled by tunable phase shifter. In the first example, a single-section device is designed to achieve the maximum possible phase range, whereas, in the second example, a phase shifter is designed to archive a full-cycle phase range, i.e., \( \Delta \Phi = 360^\circ \).

For the single-section device, the derived design equations show that a phase range equal to 255° can be achieved with more than 10-dB return loss across a 36% fractional bandwidth by using a coupled structure of length \( l = 0.083 \lambda \). The required inductor, varactor, and mode impedances are: \( L = 6.5 \text{ nH}, \ C_{\text{min}} = 0.2 \text{ pF}, R_e = 10, Z_{\text{ee}} = 10 \Omega \), and \( Z_{\text{oe}} = 228 \Omega \).

Concerning the device that has a full cycle phase range, two sections are used. Thus, the phase range required from each section is 180°. Solving (1)–(15) for 36% fractional bandwidth and the required phase range shows that several options are possible. Since one of the main objectives of the current work is to achieve the required performance using a compact and easy-to-manufacture structure, a solution that needs the smallest possible value for \( R_e \), moderate values for \( Z_{\text{oo}} \) and \( Z_{\text{oe}} \), and smallest length for the coupled structure is adopted. Moreover, the smallest value for the inductor \( L \) and the largest value for \( C_{\text{min}} \) are targeted in the solution. The reason behind targeting a small value for \( L \) and large value for \( C_{\text{min}} \) is to minimize the effect of the stray or parasitic elements in the varactor and the inductor. Concerning the varactor diode, the relative effect of the stray elements increases when using a very small value for \( C_{\text{min}} \).

For the inductor, the series resistance increases with increasing \( L \) as indicated by the technical data of microwave chip inductors. The optimum calculated values for the inductor, varactor, and mode impedances using the MATLAB code for (1)–(15) are \( L = 1.8 \text{ nH}, C_{\text{min}} = 0.5 \text{ pF}, R_e = 8, Z_{\text{oo}} = 11 \Omega \), and \( Z_{\text{oe}} = 225 \Omega \).

The proposed phase shifter is implemented in this work using parallel-coupled microstrip lines. As concluded from the design values of the two examples, high values for the even-mode impedance and low values for the odd-mode impedance are needed. The high value for \( Z_{\text{ee}} \) can be achieved by using a slotted ground plane, which results in a reduction in the even-mode capacitor and, thus, an increase in the even-mode impedance [20]. Concerning the requirement of a range of low odd-impedance values, it can be achieved by connecting a chip capacitor (\( C_{d} \)) between the middle points of the coupled lines. This capacitor has no effect on the even-mode circuit. However, it increases the equivalent odd-mode capacitor of the coupled structure and thus decreases the odd-mode impedance [21].

The final structure of a single-section tunable phase shifter is shown in Fig. 9. The top layer includes the coupled lines and the biasing circuit for the varactor diodes. The varactor diodes are connected from one side with the inductor and from the other side with the ground plane of the device located at the bottom layer. There circuit has radio frequency chokes (RFC) to isolate the microwave signal from the biasing line and dc block chip capacitors to block the biasing voltage from the input and output ports.

In order to find the initial dimensions of the coupled structure, a quasi-transverse electromagnetic propagation is assumed for the structure of Fig. 9. Thus, the even- and odd-mode impedances of the coupled lines are determined from the effective capacitances per unit length of the lines and the phase velocity in the utilized medium [17]. The complete analysis for parallel-coupled lines with or without slotted ground plane with the help of the conformal mapping technique is presented in [20]. That analysis is modified here to include the effect of the capacitor \( C_{d} \) connected between the two coupled lines as follows:

\[
Z_{\text{ee}} = \frac{60\pi}{\sqrt{\varepsilon_r}} \frac{K_1}{K_1 + K_2} \left( \frac{1}{L} \right)
\]

\[
Z_{\text{oe}} = \frac{60\pi}{\sqrt{\varepsilon_r}} \frac{K_1}{K_1 + K_2} \left( \frac{1}{L} \right)
\]

\[
k_1 = \left\{ \begin{array}{ll}
1 + \exp(-\pi(w_s / s)(2h)) & \text{if} \quad w_s < s - 2w_e / c \\
1 + \exp(-\pi(w_s / s)(2h)) & \text{if} \quad w_s > s - 2w_e / c
\end{array} \right.
\]

\[
k_2 = \frac{\tan\pi(s / (4h))}{\tan\pi((s + 2w_e) / (4h))}
\]

\[
k_3 = \frac{s}{s + 2w_e}
\]

where \( w_c \) is the width of any of the coupled lines, \( w_s \) is the width of the slot in the ground plane, and \( s \) is the width of the gap between the coupled lines (the parameters \( w_{ce}, w_{cs}, \) and \( s \) are shown in Fig. 7). \( K_1 \) and \( K_2 \) are the first kind elliptical integral and its complementary, respectively, of the parameters \( k_1, \) and \( \varepsilon_r \) and \( h \) are the dielectric constant and thickness of the substrate, respectively.

Using the analysis in [20] and (16)–(20), the initial dimensions (\( w_{ce}, w_{cs}, \) and \( s \)) for the designed devices assuming Rogers RT6010 (\( \varepsilon_r = 10.2, \ h = 0.635 \text{ mm} \)) as the substrate are calculated. The optimized values for those parameters and other design parameters calculated previously using (1)–(15) are then found using the software CST Microwave Studio. The final values for the single-section 255° phase shifter are \( w_{ce} = 0.58 \text{ mm}, \ s = 0.22 \text{ mm}, \ l = 4.9 \text{ mm}, \ w_s = 7.5 \text{ mm}, \ C_x = 1.4 \text{ pF}, \ L = 5 \text{ nH}, \ C_{\text{min}} = 0.28 \text{ pF}, \) and \( C_{\text{max}} = 2.8 \text{ pF}. \) For the full-cycle, two-section phase shifter, the values are: \( w_{ce} = 0.3 \text{ mm}, \ s = 0.22 \text{ mm}, \ l = 4.4 \text{ mm}, \ w_s = 7.5 \text{ mm}, \ C_x = 1.5 \text{ pF}, \ L = 1.7 \text{ nH}, \ C_{\text{min}} = 0.6 \text{ pF}, \) and \( C_{\text{max}} = 4.8 \text{ pF}. \)

It is well worth noting that the optimized lengths of the coupled structures as a function of the wavelength at the center frequency (2.2 GHz) are \( l = 0.085 \lambda \) and \( l = 0.076 \lambda \) for the single- and two-section devices, respectively. This result confirms the
values predicted by the theoretical model and validates the compactness of the phase shifters designed following the proposed method.

IV. RESULTS AND DISCUSSIONS

The full-wave electromagnetic simulator CST Microwave Studio is used to calculate the performance of the designed devices. For the single-section 255° phase shifter, the achieved phase range \( \Delta \Phi \) and variation of the amplitude of the \( S \)-parameters across the frequency range from 1.7 to 2.7 GHz are shown in Figs. 10 and 11. The target of 255° phase range is obviously accomplished using a coupled structure of length 0.085\( \lambda \). Across the targeted band from 1.8 to 2.6 GHz, \( \Delta \Phi \) is equal to \( 257^\circ \pm 10^\circ \). Moreover, the return loss is more than 10 dB across the whole band, whereas the insertion loss is less than 1 dB across the investigated band for all of the varactor’s capacitor values from \( C_{\text{min}} \) to \( C_{\text{max}} \).

Concerning the insertion and return losses indicated in Fig. 11, it is worth mentioning that the largest return loss and thus the smallest insertion loss are obtained at the high end of the band when the varactor’s capacitor is equal or close to \( C_{\text{min}} \). The opposite thing occurs at the low end of the band or when the varactor’s capacitor is equal or close to \( C_{\text{max}} \).

Concerning the two-section 360° phase shifter, the phase and amplitude performances of the device are shown in Figs. 12 and 13. The device achieves the required full-cycle phase range despite using coupled structures that have a total length of only 0.15\( \lambda \). Across the band from 1.7 to 2.7 GHz, \( \Delta \Phi \) is equal to \( 372^\circ \pm 20^\circ \) as revealed in Fig. 10. The return loss is more than 10 dB, whereas the insertion loss is less than 1.5 dB across the band from 1.78 to 2.55 GHz, as shown in Fig. 13.
As a final step to prove the validity of the designed phase shifter, the full-cycle device was manufactured (Fig. 14) and tested. To get the required range for \( C_r \), a GaAs hyperabrupt varactor diode with a biasing voltage range from 2 to 20 V is used. In order to minimize the effect of parastics coming from using short-circuit vias to connect the varactor diodes to the ground, those diodes were inserted in the via hole and connected directly to the ground. The biasing circuit of the diodes includes three radio frequency chokes (RFC) of 10 \( \mu \)H and two 1-nF dc block chip capacitors.

The measured phase and amplitude performances of the device are depicted in Figs. 15 and 16 for a biasing voltage changing from 2 to 20 V. The results indicate that the required full-cycle phase range \( \{360^\circ \pm 15^\circ\} \) is realized across 40% fractional bandwidth extending from 1.8 to 2.7 GHz. The return loss is more than 10 dB and the insertion loss is less than 3.2 dB across the band from 1.8 to 2.6 GHz.

The general variations in the simulated and measured phase range (Figs. 12 and 15) and amplitude of \( S \)-parameters (Figs. 13 and 16) with frequency agree well with each other. However, the measured results indicate higher insertion losses. The additional losses are believed to be due to the parasitic elements (resistor, inductor and packaging capacitor) of the varactor diodes and the effective resistor of the utilized inductors. A parametric study is performed to confirm the reasons behind the additional losses. It is found that if the parasitic elements of the diodes are assumed to have the values (inductor = 0.5 nH, resistor = 2 \( \Omega \), and capacitor = 0.1 pF), the simulated insertion loss becomes equal to the measured loss. Thus, it is expected that the parasitic elements of the utilized diodes have the above predicted values.

Table I shows a comparison between the main features of the proposed phase shifter and other reflection-type phase shifters fabricated using the printed circuit board (PCB) technology. It is clear that the presented device has the most compact size. It also has the highest fractional bandwidth with the lowest insertion loss as compared with the full-cycle PCB phase shifters. Concerning the reflection-type phase shifters designed using the
thin-film technology, the insertion loss of those devices is very high, as indicated in the table of characteristics in [11]. In that technology, amplifiers are usually used to compensate for the high insertion losses [15].

V. CONCLUSION

It has been shown that the reflection-type phase shifter can be designed using less than one tenth of a wavelength coupled structure if the mode impedances of that structure are chosen properly. The proposed method has been validated by designing two phase shifters. The first one achieves 260° phase range with more than 10-dB return loss and less than 1-dB insertion loss across 36% fractional bandwidth by using a coupled structure of 0.085λ length. The second device achieves a full-cycle phase range across the same band with more than 10-dB return loss and less than 1.5-dB insertion loss according to the simulations by using two coupled sections each has a length of 0.076λ. The measured results of a manufactured full-cycle phase shifter support the simulation results and the design approach.

REFERENCES


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