Ultra-wideband planar bandpass filters with improved stopband performances

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Ultra-Wideband Planar Bandpass Filters with Improved Stopband Performances

Sai Wai Wong

School of Electrical & Electronic Engineering

**A thesis submitted to the Nanyang Technological University in fulfillment of the
requirements for the degree of Doctor of Philosophy**

2009

DEDICATION

To

My supervisor

And

My grandmother, my parents, my brothers and Shan Shan

ACKNOWLEDGEMENTS

First of all, I would like to express my sincere thanks to the School of Electrical and Electronic Engineering of the Nanyang Technological University (NTU) for the financial support during the whole duration of my PhD studies.

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SUMMARY

The objective of this PhD thesis is to propose various structures of multi-mode resonator (MMR) for innovative exploration of ultra-wideband (UWB) bandpass filters with improved upper-stopband performances. All the UWB filters developed in this thesis can be primarily divided into the four categories: lowpass embedded UWB bandpass filter, electromagnetic bandgap (EBG) embedded MMR bandpass filter, shunt short-circuit loaded UWB bandpass filter and notch-band UWB bandpass filter. These UWB bandpass filters had been designed towards achieving the FCC-regulated UWB frequency spectrum which required good performance in the UWB passband and high/deep rejection in the out-of-band. In addition to the filter performance, size miniaturization is also the major goal of my research work. To evidently prove the theoretical results of those proposed UWB filters, all of those filters were designed and fabricated on the dielectric substrate, namely, RT/Duroid 6010 with substrate thickness 0.635 mm and loss tangent 0.0023. The measured results were all matched with the simulated results over a wide frequency range.

In addition, based on the basic idea of the proposed UWB filters, three Wilkinson power dividers with an operating band in the desired UWB range were proposed, designed and fabricated. A set of closed-form equations were derived using the even- and odd-mode approach to explain operating mechanism and execute optimal design of these power dividers. Finally, three prototypes of UWB power dividers were fabricated and measured to evaluate the predicted results.

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Chapter 1

Introduction

1.1 Motivation and Objectives

In recent years, the rapid development in fabrication technologies and the successful commercial deployment of wireless communications have been significantly improving our daily lives because we can free ourselves from the limitation of cables. The arrival of the third- and fourth-generation radio systems and the replacement of wired connections with Wi-Fi and Bluetooth enabled people to access to a wide range of information from anywhere and at any time. As the demand for higher capacity and faster service increases in wireless communication, a new technology is highly sought after in the crowded and scarce radio frequency (RF) spectrum. Since most of the interested frequency bands have been occupied by many specific usages, it has become stricter with the introduction of new technology without interfering with the existing applications. Thus, Ultra-wideband (UWB) technology offers a promising solution to revolutionize high-speed data transfers leading to new innovations and greater quality of services to end users.

As an important building block of the UWB communication systems, a UWB bandpass filter is required to have good in-band and out-of-band performance. There are several demanding features: low insertion loss over the entire ultra wide passband, sharp rejection skirts at band edges, wide and high rejection of upper stopband, high rejection of lower stopband, compact size and easy to fabricate. Usually, there are trade-offs among these desired performances. For example, the sharper rejection and wider upper-stopband can be simply achieved by cascading several sections of half- or quarter-wavelength resonators. Nevertheless, the overall circuit size will be enlarged and the insertion loss will be increased. Therefore, it is difficult to come out with an optimized design which satisfies all the desired features mentioned above. The UWB bandpass filters proposed in the literatures currently have advantages and disadvantages. The challenging task here turns to be how to design such a class of UWB bandpass filters that can realize most of the important features.

The objective of this PhD research is to carry out a systematic investigation on the resonant characteristics of various microstrip-line multiple-mode resonators (MMRs) with varied configurations and to further employ them to form a novel class of ultra-wideband bandpass filters with improved electrical performances and miniaturized size.

1.2 History and Background of UWB technology

This section gives an overview on the history of UWB technology, the important building blocks of UWB systems and the topologies of reported UWB bandpass filters. The first pulse based UWB radio was the Spark Gap radios which were first developed by Guglielmo Marconi in the late 1800's and were used to transmit Morse code later on. UWB

technology was then initially investigated in the late 1960s by Henning Harmuth from Catholic University of America, Gerald Ross and K. W. Robins from Sperry Rand Corporation [2]. From the 1960's to the 1990's, this UWB technology was mainly restricted to military and defense applications under various classified programs, such as highly secure communications [1]. The recent development of high speed microprocessors and fast switching techniques has made this UWB technology possible for commercial short-range communications. UWB systems transmit an extremely short pulse and occupy an ultra-wide frequency bandwidth. Since frequency is inversely related to time, the short-duration UWB pulses spread their energy across a wide range of frequencies with very low power spectral density. As a result, it allows the UWB signal to coexist with the current radio services. Besides the ability of sharing the frequency spectrum, UWB signals have some other advantages such as, being simpler than the direct sequence spread (DSSS) technique in providing wide bandwidth signals.

On April 22, 2002, the Federal Communications Commission (FCC) authorized the unlicensed commercial deployment of UWB technology under a strict power control in 7500 MHz spectrum [3]. Great interest in UWB communications has been aroused from both academic and industrial areas. The FCC has assigned conservative emission masks between 3.1 GHz and 10.6 GHz for commercial UWB devices – that is, -41.3 dBm/MHz or 75 nW/MHz – places them at the same level as unintentional radiators such as televisions and computer monitors. Based on the FCC regulations, UWB devices are classified into three major categories: communications, imaging, and vehicular radar. Throughout the full

range of UWB applications, applying UWB into short-range high-speed wireless communication is currently considered as one of the most promising areas.

For communication devices, the FCC has assigned different emission limits for indoor and outdoor UWB devices. The spectrum mask for outdoor devices is 10 dB lower than that for indoor devices, between 1.61 GHz and 3.1 GHz. The emission limits for indoor and outdoor UWB devices are shown in Figure 1-1 and Figure 1-2, respectively.

Figure 1-1. UWB Emission Limit for Indoor Systems.

Figure 1-2. UWB Emission Limit for outdoor Systems

According to the FCC regulations, indoor UWB devices must consist of handheld equipments, and their activities should be restricted to peer-to-peer operations inside buildings. The regulation of FCC also dictates that no fixed infrastructure can be used for UWB communication in outdoor environments. Therefore, outdoor UWB communications are restricted to handheld devices that can send information only to their associated receivers [1]. Thus, size is a key factor in UWB components and it is also one of the important objectives to miniaturize the size of the UWB filter in this thesis.

1.3 Literature Review of UWB Bandpass Filters

UWB bandpass filter is a key circuit block that regulates the frequency spectrum in UWB systems. In recent years, there have been tremendously aroused research interests in the exploration of a variety of UWB bandpass filters (BPFs) to meet the FCC-regulated

limit. Nevertheless, the conventional filter synthesis procedure is adequate only for the band pass filters with a relatively narrow passband, and is not suitable for the wideband filters. There are some new formulae for synthesizing the microstrip wideband bandpass filter such as [4], [5]. However, the application of these formulae can only achieve up to 50% fractional bandwidth. As the bandwidth increases, the accuracy will be degraded and become impractical. As the first ultra-wideband filter covering the whole UWB band in 2003, microstrip ring resonators with loaded open stubs were used to produce the dual-stopbands or attenuation poles near the lower and upper ends of the passband, i.e., 3.1 and 10.6 GHz, respectively [6]. However, this filter had several drawbacks such as unexpected passband below 3.1 GHz, narrow lower and upper stopbands; large size and complexity in configuration. In the same year [7], a microstrip line, mounted in a lossy composite substrate, made up an alternative ultra-wideband filter; however, its insertion loss was higher than 6.0 dB within the passband. In the coming sections, some typical UWB bandpass filters proposed in recent years are introduced and overviewed. They can be divided into four categories as discussed below.

1.3.1 Integration low- and high-pass filters

A wide bandpass filter can be implemented by direct cascade of the low- and high-pass filters. This approach usually exhibits good upper-stopband rejection, because the stopband of a lowpass filter is always located above the upper cut-off frequency of the UWB band. The work in [8] is one of the typical examples for this type of wideband filters. The low-pass filter can be implemented by using simple stepped-impedance structure and its synthesis approach can be found in most RF textbooks. Meanwhile, a high-pass filter was

realized by loading or attaching a few quarter guided wavelength short-ended stubs on a uniform transmission line. The bandpass filter can be constructed by directly cascading these low- and high-pass filters. In order to reduce the size, the composite bandpass filter was constructed by mixing the low- and high-pass filters to reduce the total circuitry size whereas keeping the performances almost the same. The lower and higher cutoff frequencies of the bandpass filter were decided by those of the high- and low-pass filters, respectively. The proposed filter has a wide stopband due to the effect of the low-pass filter. The upper and lower 3-dB cut-off frequency was designed at 3.1 GHz and 10.0 GHz to meet the requirement of the UWB system with a rejection upper-stopband up to 20.0 GHz.

The UWB bandpass filter was designed with a good out-of-band performance in [9] on a basis of the similar theory. The wide bandwidth was constructed by loading the short-circuited stubs in shunt on a uniform transmission line; however, the filter processed periodic response due to its distributive nature. To suppress the harmonic passbands, electromagnetic bandgaps implemented by capacitive loaded transmission lines were placed in between the short-circuited stubs. The proposed filter therefore achieved a desired wide passband with widened upper-stopband. Although the idea to embed the high-pass filter (low-pass filter) into low-pass filter (high-pass filter) reduced the overall circuit size greatly, the circuit size was still considerably large and the short-ended stubs made the fabrication complicated.

Short-ended stubs in [8] and [9] involve via holes. Filter with coplanar waveguide can avoid using via holes, because the ground plane and strip conductor are located on the same interface of the printed circuit board (PCB). In [10], a wideband coplanar-waveguide (CPW)

bandpass filter was achieved by cascading three-stage CPW lowpass filters and two-stage highpass filter. Since multi-stage was used, therefore the upper and lower roll-off were very sharp. However, the drawback was very large circuit size. Similar idea was implemented in [11] with size reduction by cascading single-stage CPW lowpass filter and single-stage CPW high pass filter. In this design, the fractional bandwidth was increased from 80% in [10] to 145% in [11]. Alternatively, suspended strip line in [12] could achieve very low loss in the realized passband since it was shielded by metal case. Different from [10] and [11], circuit in [12] was fabricated on two side PCB, consisting of CPW lowpass filter at the top and CPW high-pass filter at the bottom. This filter had a very wide passband ranging from 2.6 GHz to 11.0 GHz. Nevertheless, the drawback was very high cost, because the metal case and two layers of CPW circuits were needed.

1.3.2 UWB filter with multiple-mode resonator

Figure 1-3. Layout of the MMR UWB bandpass filter proposed in [16]

The concept of multiple-mode resonator (MMR) was firstly proposed in [13] that the first two resonant modes of the constituted MMR could be utilized together with the input/output parallel-coupled lines to achieve a 60% to 80% wide passband with four

transmission poles. In [14] and [15], the first three resonant modes of an improved MMR were newly constructed to realize five transmission poles with lowered insertion loss in the whole passband. Later on, the UWB bandpass filters based on this MMR concept were designed by using the first three resonant modes of this MMR together with strong coupling at the input and output ports [16]-[20]. In [16], the MMR was composed by a half wavelength low characteristic impedance section at the center and two quarter wavelength high characteristic impedance sections at two sides. The schematic of this MMR UWB bandpass filter and its performance are shown in Figure 1-3 and Figure 1-4, respectively. The quarter wavelength coupling section with respect to UWB center frequency, i.e. 6.85 GHz, first constructed a passband at the required coupling peak at 6.85 GHz. The three resonant modes of MMR within the UWB band then widened the passband, and its first and third resonant modes could be used to control the upper and lower cut-off frequency of UWB passband. As a result, a good UWB bandpass filter was constructed as shown in Figure 1-4.

With the same concept, backside aperture could be used in the quarter wavelength coupling section to enhance the coupling degree [17]. Thus this MMR based filter could be fabricated on a thin substrate (0.635 mm) and the overall size could be reduced. Furthermore, UWB bandpass filter with hybrid microstrip/CPW structure [18] and short-circuited CPW [19] also exhibited good UWB passband performance. Nevertheless, all the filters in [13]-[19] suffered from a problem of narrow upper-stopband, thus capacitive-ended interdigital coupled-lines were proposed in [20] to widen the upper-stopband. In [21], the upper-stopband could be further widened by using open-stub MMR and the size of the

filter was also reduced. In addition, the coupling gap could then be increased from 0.05 mm in [16] to 0.1 mm in [21], so that the circuit was easier for fabrication due to relaxed tolerance.

Figure 1-4. Simulated and measured results for the UWB bandpass filter proposed in [16]

1.3.3 Open- and short-circuit stub loading

Since the FCC strictly regulates the spectrum mask for the UWB applications, the sharp roll-off at the upper and lower edges is a desired feature for UWB bandpass filters. Normally, this sharp rejection at the lower and upper edges could be realized by simply cascading more stages of the UWB bandpass filters in [6]-[21], but this would dramatically increase the circuit size. One way to realize the sharp rejection skirt at two edges was to introduce two transmission zeros at both lower and upper stopbands near the passband ends. In [22], the open-ended stub was loaded at the center of the line resonator and it brought out such two transmission zeros at the lower and upper edges, respectively. However, the

bandwidth (2.2–4.4 GHz) was not wide enough to meet the UWB requirement. In order to realize requested ultra-wide bandwidth, the two transmission zeros brought by the loading stub should be at least three times apart i.e. $10.6/3.1 \approx 3.4$. Similar idea was applied in the bandpass filter presented in [23]. Again, as a main drawback, the fractional bandwidth of these constituted bandpass filters were still narrower than the specified one. A wideband synthesizing method was proposed in [24] by using Z-transform technique. Three sets of shunt short-circuit stub parameters were first calculated for the wideband filter. The short-circuit stub was then replaced by two-section open-circuit stubs. Finally, a wideband bandpass filter with transmission zeros at lower and upper-stopband was designed and fabricated with 84% fractional bandwidth. However, it was not DC blocking, therefore there was an unwanted passband at low frequency.

1.3.4 Other UWB bandpass filters

Besides the main topologies of various UWB bandpass filters discussed above, there were also other methods for constructing UWB bandpass filters as well as the wide-band bandpass filters [25]-[32]. One of the main difficulties in the design of an UWB bandpass filter was the 110% ultra-wide fractional bandwidth. In this way, coupling elements were of critical importance to achieve ultra-wide bandwidth. In order to have very tight coupling, [25] utilized the broadside coupling between microstrip to coplanar waveguide. The open-end CPW section was fabricated on the ground plane, and two microstrip lines separated with a gap were fabricated on the top plane. Therefore, the configuration of this filter became compact and the broadside coupling structure through dielectric substrate had very tight coupling strength. In [26], the etched slots in the ground plane of the parallel coupled-

lines could significantly increase the coupling degree, resulting in a 100% fractional bandwidth. The resonant frequency spectrum of the SIR was investigated in [27] and broad-band quasi-Cheyshev bandpass filters were developed based on the multi-mode SIR. Dual-Mode ring resonator was used to design a wide-band bandpass filter with improved stopband property in [28], and exhibiting a 50% fractional bandwidth with compact size.

Most of the UWB bandpass filters discussed in [4]-[32] exhibited very good performance and they were suitable for practical implementation. Nevertheless, existing radio signals such as wireless local-area network (WLAN) at around 5.6 GHz may interfere with the UWB radio system within the defined UWB frequency spectrum. Thus, a narrow band notched UWB bandpass filter has been strongly demanded in practical implementation. The authors of [25] realized a dual-band UWB bandpass filter based on their original work by integrating a stub in the broadside-coupled conductors in their later work [33]. The resonance of the stub introduced a new rejection band inside the UWB passband which then resulted in a dual-band filtering. The rejection band could be adjusted to some specific band by adjusting the length of the stub. In [34], the notch band was realized based on the initial UWB bandpass filter in [32]. This notched band was created by adding open stubs in first and last connecting lines. Another example of band notched filter was based on the asymmetric parallel-coupled lines [35], where the UWB passband could be first constituted by the filter in [20]. The interdigital parallel-coupled lines were then modified to be asymmetrically reformatted to create a narrow notch band within the UWB passband. The frequency of the notch band could be controlled by the length difference between the two coupling lines.

1.4 Achievements of the Research

The author's major achievements made so far are summarized in the following:

- 1) The synthesis approach of UWB filter is a challenging work, due to its wideband performance. The existing synthesis approaches are based on the assumption of narrowband filter. In this thesis, a synthesis approach based on Chebyshev function with particular MMR structure was developed to directly extract the characteristic impedance of the MMR, and the even and odd mode characteristic impedances of the coupled-line. These extracted values can be converted to physical dimensions of the filter. As a result, it provides a guideline for designing UWB filters with different MMR structures.
- 2) Most of the reported UWB filters had very good passband. However, most of them suffer poor upper-stopband performance. The author proposed several MMR structures to solve this problem. These MMR structures were investigated and embedded to replace the half-wavelength resonator in [16], the overall size of the resultant UWB bandpass filter becomes much compact with high attenuation in the upper unwanted passband, i.e., with wide/deep upper stopband [36]-[40].
- 3) In addition to good upper-stopband, a sharp rejection skirt is very important for UWB filter to meet the FCC regulation. The proposed short-circuit stub-loaded MMR [43] was able to improve the rejection skirt. By observing the chart for the mode distribution, a wide upper-stopband can be achieved without size-increased multi-stage approaches in [36]-[40].

- 4) Since UWB covers a very wide bandwidth, e.g. 3.1-10.6 GHz. Its frequency band may overlap with some existing wireless systems and causes interference with these existing wireless systems. In this thesis, two kinds of notch-band filters were proposed to avoid the interference with existing radio systems such as WLAN [41], [42].
- 5) Based on the concept of the UWB filters developed in this thesis, three Wilkinson power dividers were designed to realize the good in-band performances in the specified UWB passband with improved out-of-band improvement. Only one isolation resistor was required for good isolation between two output ports in the entire UWB [44]. Closed form formulae were derived and design curves were plotted to provide useful guidelines to design UWB power divider.
- 6) For the industry point of view, filters are required to be low cost, compact and easy for fabricate. All the filters and power dividers in this thesis satisfied those requirements as they were designed based on microstrip line on a single layer PCB board.

1.5 Organization of the Dissertation

Chapter 1 is an introductory chapter that briefly describes history and background of various UWB technologies developed so far. Besides, a literature review of various UWB bandpass filters is carried out to distinguish their advantages and disadvantages.

Chapter 2 describes a systematic theory for synthesis design of traditional UWB bandpass filters in terms of Chebyshev transfer function. Two design examples are given in this chapter to explain how this technique can be limitedly applied to these UWB filters.

Chapter 3 presents several UWB bandpass filters with improved upper-stopband performances by cascading lowpass filters in the two input sides of an initial UWB filter. A lowpass filter is then embedded into the MMR to miniaturize the circuit size of the resultant UWB filter with the same properties. Two-stage MMR sections are cascaded to further enhance the concerned rejection skirts in the out-of-band.

Chapter 4 introduces an EBG-embedded MMR to construct an alternative UWB bandpass filter with improved upper-stopband performances. As the two EBG unit-cell structures are studied in terms of their attenuation and propagation constants to show their slow-wave properties, they are embedded into the resonator to achieve good UWB passband, highly attenuated upper-stopband and compact size. Again, two-stage EBG-embedded MMR is utilized to further improve the rejection skirts in their out-of-bands performance.

Chapter 5 provides another approach to improve the rejection skirts with single MMR resonator except cascading MMRs. This can be achieved by adding shunt short-circuit stubs in the resonator. The filter in [16] is modified by tap-connecting the coupled-line to the half-wavelength resonance section. As the mode chart is plotted and studied, some additional transmission zeros can be generated and properly allocated such that a desired wide upper-stopband can be created. Furthermore, two short-circuit stubs are mounted on the MMR so as to further improve the rejection skirts.

Chapter 6 is the extension of chapter 3 and 4. The UWB filters proposed in chapter 3 and 4 are modified in their configurations to create a notch-band within the passband of UWB. Two approaches are developed using either asymmetric coupled arms or asymmetric shunt stubs. Of them, the latter one not only produces a controllable notch-band, but also sharpens the rejection skirts as inquired.

Chapter 7 proposes three UWB power dividers with a single isolation resistor that has good power dividing performance in the UWB passband and good out-of-band rejection. The structures reported in chapter 3-6 are helpful in constructing these power dividers. The even- and odd-mode analysis approach is described to analysis and design of these power dividers. Next, some prototypes of these dividers are fabricated and the measured results confirmed well with the predicted results.

Chapter 8 is the conclusion and recommendations for this PhD thesis.

Chapter 2

Synthesis Technique of UWB Filter

2.1 Fundamental Elements of UWB Filter

Generally speaking, a microwave filter consists of a number of fundamental elements such as inductors, capacitors, resonators, coupling elements and so on. In the design of UWB bandpass filters in this thesis, microstrip resonator with multiple resonances and parallel-coupled microstrip line were two primary building elements. Their performances need to be extensively characterized using the fundamental transmission-line theory.

2.1.1 Fundamental theory of microstrip resonators

Microwave resonators have been utilized for a variety of applications, including filters, oscillators, frequency meters and tuned amplifiers. Herein, I will focus on the application of resonators in wideband microstrip line filters. Microstrip line resonators are commonly used in microwave filter designs for wireless communication systems. Although these resonators do not have high Q value compared to waveguide and dielectric resonators, they have valuable features such as simple structure, low cost, small size and the capability of

integrating with active circuits. Figure 2-1 shows the fundamental structure of a microstrip line half-wavelength resonator with two open-circuited ends, which is a typical transmission line resonator. The conducting strip is on the top of a dielectric substrate and the ground conductor is at the bottom. The work in [16] gives an example for this uniform half-wavelength resonator for UWB filter as discussed in previous chapter.

Figure 2-1. Fundamental structure of a microstrip line half-wavelength resonator

The input admittance Y_{in} at one opened-end, looking toward its opposite opened-end, of such a transmission line resonator can be expressed in the following,

$$Y_{in} = jY_0 \tan \beta l$$

At the opened-end, $Y_{in} = 0$, thus deriving the resonant condition valid for all the resonant modes.

$$\beta l = n\pi$$

where $n = 1, 2, 3, \dots$

Thus the resonant frequencies can be expressed as follows,

$$f = \frac{cn}{2l\sqrt{\epsilon_{\text{reff}}}}$$

where c is the velocity of light in free space, l is the length of the resonator and ϵ_{reff} is the relative permittivity of this inhomogeneous microstrip line. All the resonant modes in this resonator can be classified into two distinct modes, i.e., odd-order modes ($n = 1, 3, 5 \dots$) and even-order modes ($n = 2, 4, 6 \dots$). As mentioned earlier, the center plane of the resonator presents a perfect electric wall (P.E.W.) or short circuit for the odd-order modes and a perfect magnetic (P.M.W.) wall or open circuit for the even-order modes as depicted in Figure 2-1. Therefore, loading an open-circuited stub in shunt at the center position does not affect the odd-order modes but influences the even-order modes greatly.

Microstrip resonators may have varied configurations for different applications. Stepped impedance resonator (SIR) is a widely used non-uniform resonator which provides more degree of freedom for controlling the resonant frequencies. Figure 2-2 shows the typical structure of an SIR. Similar to the analysis of the uniform line resonator, resonant performances of SIR can also be characterized by using the well-known even- and odd-mode approach [30], [27]. According to the resonant conditions, a set of equations can be derived as below:

$$\tan \theta_1 = R \cot \theta_2 \quad (\text{odd-modes})$$

$$\cot \theta_1 = -R \cot \theta_2 \quad (\text{even-modes})$$

where $R = Z_2/Z_1$ is the impedance ratio of the two dissimilar sections in the SIR.

For a given R value, one can use a root finding program to determine its multiple resonant frequencies. Other configurations of resonators for UWB filters are proposed and discussed in the following chapters.

Figure 2-2. Typical structure of a microstrip line stepped impedance resonator (SIR).

2.1.2 Theory of microstrip parallel coupled-line

When two uniform microstrip lines are placed in close proximity to each other as shown in Figure 2-3, a fraction of the power on the main line can be coupled to the secondary line. Coupled-line can be of any form depending on the applications [46], [47]. The structure shown in Figure 2-3 is called parallel-coupled line structure via coupling between two edges of the coupled strip conductors on the top interface.

Figure 2-3. Structure of parallel microstrip coupled-line.

In general, a transmission line with $n+1$ strip conductors in an inhomogeneous dielectric supports n propagating transverse electromagnetic (TEM) modes with distinct phase velocity and characteristic impedance. Parallel microstrip coupled-line can be analyzed using even- and odd-mode approach to calculate the different phase velocities for dominant even- and odd-mode [48]. In a more general case, coupled-lines with asymmetric strip widths need to be characterized in terms of c-mode and π -mode as discussed in [49]. To simplify the analysis, these phase velocities are approximately considered as the same value and this assumption has been widely used in many other works at low frequencies. Based on the transmission line theory, the ABCD-matrix can be expressed as

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \frac{Z_+ \cos(\theta)}{Z_-} & j \frac{Z_-^2 - Z_+^2 \cos^2(\theta)}{2Z_- \sin(\theta)} \\ j \frac{2 \sin(\theta)}{Z_-} & \frac{Z_+ \cos(\theta)}{Z_-} \end{bmatrix} \quad (2.1)$$

where $Z_+ = Z_{oe} + Z_{oo}$ and $Z_- = Z_{oe} - Z_{oo}$. θ is the electrical length of the coupled-line. Z_{oe} and Z_{oo} correspond to even- and odd-mode impedances of the coupled-line respectively.

2.1.3 Theory of microstrip three-line coupled-line

As discussed in previous section, three-line microstrip coupled-line consists of three strip conductors and one common ground plane as shown in Figure 2-4. It has three distinct modes, namely, a-, b- and c-mode [50]-[54]. If the three lines are equal or symmetric in their widths and all their phase velocities are equal, the ABCD-matrix of this two-port interdigital three-line microstrip coupled-line can be derived as

$$A = \frac{(Z_{b2}R_{b2} - Z_{c2}R_{c2})\cos(\theta)}{Z_{b2} - Z_{c2}} \quad (2.2a)$$

$$B = j \frac{(Z_{b2}R_{b2} - Z_{c2}R_{c2})(Z_{b2}R_{c2} - Z_{c1}R_{b2})\cos^2(\theta) - 2(Z_{b2} - Z_{c2})}{2(R_{b2} - R_{c2})\sin(\theta)} \quad (2.2b)$$

$$C = \frac{j(R_{b2} - R_{c2})\sin(\theta)}{Z_{b2} - Z_{c2}} \quad (2.2c)$$

$$D = \frac{Z_{b2}R_{c2} - Z_{c1}R_{b2}\cos(\theta)}{2(Z_{b2} - Z_{c2})} \quad (2.2d)$$

where Z_{jk} , R_{jk} are the characteristic impedance and mode-voltage ratio of the b-mode to the a-mode respectively, of line k for j -mode [52]. θ is the electrical length of the coupled-line.

Figure 2-4. Structure of interdigital three-line microstrip coupled-line.

Figure 2-5. S_{21} -parameters of the three-line and two-line coupled-line.

Figure 2-5 plots the S_{21} -parameters of the three-line and parallel coupled-line performance with same strip width and strip gap. They exhibit a good wideband performance. Nevertheless, three-line microstrip coupled-line shows a wider passband than the parallel coupled-line due to enhanced coupling degree among three lines. Thus three-line microstrip coupled-line is a better candidate for designing UWB bandpass filter.

2.2 UWB Synthesis Approach by Chebyshev Function

Traditional filter synthesis method is based on assumption of narrow band performance and those approaches can only valid up to 30% fractional bandwidth [55]-[57]. For the UWB bandpass filter with 110% fractional bandwidth, the synthesis method is under developing in the recent years [58]-[60]. By using transmission line theory and introducing the Chebyshev function as a transfer function, one can develop the synthesis method for certain UWB bandpass filters with simple and regular configurations. In the following, two of these filters are discussed and designed via this approach

2.2.1 Transfer function of UWB filter with uniform resonator

Figure 2-6. Schematic of UWB filter for half-wavelength resonator.

Figure 2-6 shows the schematic of UWB filter proposed in [16] which exhibits a very good UWB passband. This filter consists of two coupled-line sections at both sides and one low-impedance transmission line at the center. The former have been analyzed in section 2.1.2 and their performance can be expressed using ABCD-matrix T_c , and the latter can be modeled using ABCD-matrix, T_u as below,

$$T_u = \begin{bmatrix} \cos(2\theta) & jZ_u \sin(2\theta) \\ jY_u \sin(2\theta) & \cos(2\theta) \end{bmatrix} \quad (2-3)$$

 Synthesis Technique of UWB Filter

where Z_u is the characteristic impedance of half-wavelength resonator. By using the Richard's transformation [61] $S=j\tan(\theta)$, T_c and T_u can be written as:

$$T_c = \frac{1}{Z_- \sqrt{1+S^2}} \begin{bmatrix} Z_+ & j \frac{Z_-^2(1+S^2) - Z_+^2}{2S} \\ j2S & Z_+ \end{bmatrix} \quad (2-4)$$

$$T_u = \frac{1}{1+S^2} \begin{bmatrix} 1-S^2 & j2Z_u S \\ j2S/Z_u & 1-S^2 \end{bmatrix} \quad (2-5)$$

From (2-4) and (2-5), the overall ABCD-matrix of the UWB filter in Figure 2-6 can be derived by

$$T \begin{bmatrix} A & B \\ C & D \end{bmatrix} = T_c \cdot T_u \cdot T_c \quad (2-6)$$

where

$$A = \frac{1}{Z_u} (2Z_+^2 Z_u - 2Z_+^2 Z_u S^2 - Z_+ Z_-^2 - Z_+ Z_-^2 S^2 + Z_+^3 - 4Z_u^2 Z_+ S^2 - Z_u Z_-^2 + Z_u Z_+^2 S^4) \quad (2-6a)$$

$$B = \frac{-j}{2Z_u S} (-2Z_+ Z_u Z_-^2 + 2Z_+^3 Z_u + 2Z_+ Z_u Z_-^2 S^4 - 2Z_+^3 Z_u S^2 + Z_-^4 + 2Z_-^4 S^2 - 2Z_+^2 Z_-^2 + Z_-^4 S^4 - 2Z_+^2 Z_-^2 S^2 + Z_+^4 - 4Z_+^2 Z_u^2 S^2) \quad (2-6b)$$

$$C = \frac{2jS}{Z_u} (2Z_+ Z_u - 2Z_+ Z_u S^2 + Z_+^2 - 4Z_u S^2) \quad (2-6c)$$

$$D = \frac{1}{Z_u} (2Z_+^2 Z_u - 2Z_+^2 Z_u S^2 - Z_+ Z_-^2 - Z_+ Z_-^2 S^2 + Z_+^3 - 4Z_u^2 Z_+ S^2 - Z_u Z_-^2 + Z_u Z_+^2 S^4) \quad (2-6d)$$

where $Z_+ = Z_{0e} + Z_{0o}$ and $Z_- = Z_{0e} - Z_{0o}$ and four elements (A, B, C, D) of this ABCD-matrix, T , can be calculated from (2-6) and they can be simply converted to their corresponding S-matrix parameter [57]. After some tedious mathematical derivation is done, the power transmission coefficient, $|S_{21t}|^2$, of this UWB filter can be deduced as:

$$|S_{21t}|^2 = \frac{16Z_-^4 Z_u^2 Z_0^2 (1+S^2)^4 S^2}{a_0' + a_2' S^2 + a_4' S^4 + a_6' S^6 + a_8' S^8 + 16Z_-^4 Z_u^2 Z_0^2 S^{10}} \quad (2.7)$$

$$= \frac{(1+S^2)^4 S^2}{\frac{1}{16Z_-^4 Z_u^2 Z_0^2} (a_0' + a_2' S^2 + a_4' S^4 + a_6' S^6 + a_8' S^8) + S^{10}}$$

where the coefficient of a_i' is a function of Z_+ , Z_- and Z_u ($i=0,2,4,6,8$)

$$a_0' = 4Z_+^7 Z_u - 4Z_+^2 Z_-^6 + 6Z_+^4 Z_-^4 + 4Z_+^2 Z_-^4 Z_u^2 - 8Z_+^4 Z_-^2 Z_u^2 + 4Z_+^6 Z_u^2 - 4Z_+^6 Z_-^2 + Z_+^8 + 12Z_+^3 Z_-^4 Z_u - 4Z_+ Z_-^6 Z_u - 12Z_+^5 Z_-^2 Z_u$$

$$a_2' = -4Z_+^6 Z_-^2 - 4Z_+^7 Z_u + 12Z_+^4 Z_-^4 - 16Z_+^6 Z_u^2 - 12Z_+^2 Z_-^6 + 8Z_+^6 Z_0^2 - 16Z_+^5 Z_u^3 + Z_+^8 + 4Z_-^8 - 8Z_+^2 Z_-^4 Z_u^2 - 8Z_+ Z_-^6 Z_u + 24Z_+^4 Z_-^2 Z_u^2 + 16Z_+^3 Z_-^2 Z_u^3 + 16Z_-^4 Z_u^2 Z_0^2 + 12Z_+^3 Z_-^4 Z_u + 32Z_+^5 Z_u Z_0^2 - 16Z_+^4 Z_-^2 Z_0^2 - 32Z_+^2 Z_-^2 Z_u^2 Z_0^2 - 48Z_+^3 Z_-^2 Z_u Z_0^2 + 16Z_+ Z_-^4 Z_u Z_0^2 + 8Z_+^2 Z_-^4 Z_0^2 + 32Z_+^4 Z_u^2 Z_0^2$$

$$a_4' = 16Z_+^4 Z_u^4 + 16Z_+^4 Z_0^4 + 6Z_+^4 Z_-^4 - 12Z_+^2 Z_-^6 + 4Z_+^6 Z_u^2 + 16Z_+^5 Z_u^3 + 64Z_+^3 Z_u Z_0^4 + 64Z_+^2 Z_u^2 Z_0^4 + 32Z_-^4 Z_u^2 Z_0^2 + 12Z_+^5 Z_-^2 Z_u - 24Z_+^2 Z_-^4 Z_u^2 - 12Z_+^3 Z_-^4 Z_u + 24Z_+^4 Z_-^2 Z_u^2 - 16Z_+^4 Z_-^2 Z_0^2 + 96Z_+^2 Z_-^2 Z_u^2 Z_0^2 - 128Z_+^3 Z_u^3 Z_0^2 + 64Z_+ Z_-^2 Z_u^3 Z_0^2 + 6Z_-^8 - 32Z_+^5 Z_u Z_0^2 - 128Z_+^4 Z_u^2 Z_0^2 + 16Z_+^2 Z_-^4 Z_0^2 + 16Z_+ Z_-^4 Z_u Z_0^2$$

$$a_6' = -4Z_+^2 Z_-^6 - 256Z_+ Z_u^3 Z_0^4 - 64Z_+^3 Z_u Z_0^4 + 8Z_+^2 Z_-^4 Z_u^2 - 256Z_+^2 Z_u^2 Z_0^4 - 8Z_+^4 Z_-^2 Z_u^2 + 4Z_-^8 + 8Z_+ Z_-^6 Z_u - 12Z_+^3 Z_-^4 Z_u + 32Z_-^4 Z_u^2 Z_0^2 - 16Z_+^3 Z_-^2 Z_u^3 + 96Z_+^2 Z_-^2 Z_u^2 Z_0^2 + Z_+^2 Z_-^4 Z_0^2 + 48Z_+^2 Z_-^3 Z_u Z_0^2 + 32Z_+^4 Z_u^2 Z_0^2 + 128Z_+^2 Z_u^4 Z_0^2 + 128Z_+^3 Z_u^3 Z_0^2 - 16Z_+ Z_-^4 Z_0^2$$

$$a_8' = 256Z_u^4 Z_0^4 + 64Z_+^2 Z_u^2 Z_0^4 + 256Z_+ Z_u^3 Z_0^4 + 8Z_-^8 + 32Z_-^4 Z_u^2 Z_0^2 + 4Z_+ Z_-^6 Z_u + 4Z_+^2 Z_-^4 Z_u^2 - 64Z_+ Z_-^2 Z_u^3 Z_0^2 - 16Z_+ Z_-^4 Z_u Z_0^2 - 32Z_+^2 Z_-^2 Z_u^2 Z_0^2$$

$$a_0 = \frac{a'_0}{16Z_-^4 Z_u^2 Z_0^2}, a_2 = \frac{a'_2}{16Z_-^4 Z_u^2 Z_0^2}, a_4 = \frac{a'_4}{16Z_-^4 Z_u^2 Z_0^2}, a_6 = \frac{a'_6}{16Z_-^4 Z_u^2 Z_0^2}, a_8 = \frac{a'_8}{16Z_-^4 Z_u^2 Z_0^2}$$

2.2.2 Chebyshev transfer function for synthesis of UWB filter

The transfer function of the UWB filter can be expressed in terms of Chebyshev function $H_N(\theta)$,

$$|S_{21c}|^2 = \frac{1}{1 + r^2 |H_N(\theta)|^2} \quad (2.8)$$

where r is the ripple constant related to a given passband ripple L_{Ar} in dB by

$$r = \sqrt{10^{L_{Ar}/10} - 1}$$

and $H_N(\theta)$ is defined as the Chebyshev function:

$$H_N(\theta) = \frac{T_N(\cos(\theta)/\cos(\theta_c))}{d \sin(\theta)} \quad (2.9)$$

$T_N(x) = \cos(N \cos^{-1} x)$ is the Chebyshev function of first kind and order of degree N . θ_c is the electric length at lower cutoff frequency. θ is the electric length that satisfies $\theta = \theta_c \times f/f_c$. d is the constant factor to normalize the function $H_N(\theta)$, so that it satisfies $0 \leq |H_N(\theta)|^2 \leq 1$ in passband. The function $H_N(\theta)$ in (2-9) is difficult to solve, thus the $H_N(\theta)$ is expressed by the Chebyshev polynomial approximation in (2-10) by choosing $N=5$.

$$H_5(\theta) = \frac{1}{d \sin(\theta)} \left(\frac{16 \cos^5 \theta}{\cos^5 \theta_c} - 20 \frac{\cos^3 \theta}{\cos^3 \theta_c} + 5 \frac{\cos \theta}{\cos \theta_c} \right) \quad (2.10)$$

where the normalized factor is calculated as $d=1.105$ for (2-10). After Richard's transformation of (2-10) and substitute into (2-8), the transfer function can be derived in (2-11) which is proven to have the same form in polynomials with regard to Richard's variable, $S=\tan(\theta)$

$$|S_{21c}|^2 = \frac{(1+S^2)^4 S^2}{b_0 + b_2 S^2 + b_4 S^4 + b_6 S^6 + b_8 S^8 + S^{10}} \quad (2.11)$$

$$b_0 = (r^2 + 25r^2 S_c^2 + 200r^2 S_c^4 + 560r^2 S_c^6 + 640r^2 S_c^8 + 256r^2 S_c^{10}) / d^2$$

$$b_2 = (d^2 - 20r^2 - 300r^2 S_c^2 - 1080r^2 S_c^4 - 1440r^2 S_c^6 - 640r^2 S_c^8) / d^2$$

$$b_4 = (110r^2 + 4d^2 + 630r^2 S_c^2 + 1080r^2 S_c^4 + 560r^2 S_c^6) / d^2$$

$$b_6 = (6d^2 - 100r^2 - 300r^2 S_c^2 - 200r^2 S_c^4) / d^2$$

$$b_8 = (25r^2 + 4d^2 + 25r^2 S_c^2) / d^2$$

Figure 2-7. Calculated S_{21} -and S_{11} -parameters of the Chebyshev transfer function for the half-wavelength resonator UWB filter.

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Equation (2-8) with $N=5$ can be plotted in Figure 2-7 with cutoff frequency $\theta_c=45^\circ$ and ripple value $r=0.2$ ($L_{Ar}=15.37$ dB). The graph exhibits a very good UWB performance with almost equal ripples and five transmission poles in the passband that are very close to the result in [58]. If the transfer function in (2-7) is enforced to have similar performance with Chebyshev transfer function in Figure 2-7, the coefficients a_i must be equal to b_i ($i=0,2,3,4,6$). Thus by solving the unknown parameters of a_i , one can obtain layout dimensions of the UWB filter. In this case, three variables Z_{oe} , Z_{oo} and Z_u have to be determined. The first three equations can be utilized to solve these three unknowns so that the passband response of this circuit can be same as in Figure 2-7. By solving the system equations in Matlab, $Z_{oe}=148.1420\ \Omega$, $Z_{oo}=35.7463\ \Omega$ and $Z_u=41.0649\ \Omega$ can be obtained. These values can be converted to physical circuit parameters easily by using LinCalc in ADS. The optimized circuit layout in EM simulation is shown in Figure 2-8. Since the calculated Z_{oe} , Z_{oo} and Z_u are ideal and lossless, the optimized physical dimensions shown in Figure 2-8 are slightly different from the calculated parameters.

Figure 2-8. The optimized layout dimensions for the EM simulation

Figure 2-9. The S_{21} - and S_{11} -parameters of circuit and EM simulation

The calculated circuit parameters are then utilized as the design parameters of the circuit simulation in ADS. The $|S_{11}|$ and $|S_{21}|$ can be plotted in Figure 2-9 (solid lines). The passband is almost the same as that in Figure 2-7 with the return loss around 15.3 dB. The layout of the UWB filter in Figure 2-8 is optimized by Full-wave simulation in Momentum and its $|S_{11}|$ and $|S_{21}|$ is also plotted in Figure 2-9 (dotted lines). As it can be observed in Figure 2-9, the EM simulation result (dotted lines) is matched with the circuit simulation results (solid lines). Since the loss tangent is 0.0023, conductor loss of copper and frequency dispersion of microstrip line are considered in the EM simulation, so there are discrepancies such as deteriorated return loss and narrower bandwidth when compared with the circuit simulation (ideal case).

2.2.3 Transfer function of UWB filter with short-circuit stub resonator

In previous section, a synthesis method of UWB filter is presented with uniform half-wavelength resonator. This approach can be applied in other shape of resonator, e.g. shunt short-circuit stub resonator as shown in Figure 2-10. The ABCD-matrix of coupled-line section T_c is the same as that in the previous example, but the T_u is modified to a shunt short-circuit stub after Richard's transformation as shown in (2-12).

$$T_u = \begin{bmatrix} 1 & 0 \\ 1/(jZ_u \tan(\theta)) & 1 \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ -j/(Z_u S) & 1 \end{bmatrix} \quad (2-12)$$

Figure 2-10. Schematic of UWB filter for half-wavelength resonator.

The ABCD-matrix of short-circuit stub has simpler expression than the half-wavelength resonator, thus the transfer function can also be simplified as shown in (2-13).

$$\begin{aligned} |S_{21r}|^2 &= \frac{64Z_-^4 Z_u^2 Z_0^2 (1+S^2)^2 S^6}{a_0' + a_2' S^2 + a_4' S^4 + a_6' S^6 + a_8' S^8 + 64Z_-^4 Z_u^2 Z_0^2 S^{10}} \\ &= \frac{(1+S^2)^2 S^6}{\frac{1}{64Z_-^4 Z_u^2 Z_0^2} (a_0' + a_2' S^2 + a_4' S^4 + a_6' S^6 + a_8' S^8) + S^{10}} \end{aligned} \quad (2-13)$$

$$a_0' = Z_+^8 + Z_-^8 + 6Z_+^4 Z_-^4 - 4Z_+^2 Z_-^6 - 4Z_+^6 Z_-^2$$

$$a_2' = 24Z_+^5 Z_-^2 Z_u + 4Z_-^8 + 8Z_+ Z_-^6 Z_u - 24Z_+^3 Z_+^4 Z_u - 12Z_+^2 Z_-^6 + 8Z_+^6 Z_0^2 \\ + 12Z_+^4 Z_-^4 - 8Z_+^7 Z_u - 4Z_+^6 Z_-^2 + 8Z_+^2 Z_-^4 Z_0^2 - 16Z_+^4 Z_-^2 Z_0^2$$

$$a_4' = 6Z_+^8 - 32Z_+ Z_-^4 Z_u Z_0^2 + 16Z_+^6 Z_u^2 + 6Z_+^4 Z_-^4 - 32Z_+^4 Z_-^2 Z_u^2 + 24Z_+^5 Z_-^2 Z_u + 16Z_+^2 Z_-^4 Z_0^2 - 12Z_+^2 Z_-^6 \\ + 48Z_+^3 Z_-^4 Z_u + 16Z_+^4 Z_0^2 - 16Z_+^4 Z_-^2 Z_0^2 + 96Z_+^3 Z_-^2 Z_u Z_0^2 + 24Z_+ Z_-^6 Z_u + 16Z_+^2 Z_-^4 Z_u^2 - 64Z_+^5 Z_u Z_0^2$$

$$a_6' = 4Z_-^8 - 64Z_+ Z_-^4 Z_u Z_0^2 - 24Z_+^3 Z_-^4 Z_u - 32Z_+^4 Z_-^2 Z_u^2 + 96Z_+^3 Z_-^2 Z_u Z_0^2 + 8Z_+^2 Z_-^4 Z_0^2 - 4Z_+^2 Z_-^6 \\ + 64Z_-^4 Z_u^2 Z_0^2 + 32Z_+^2 Z_-^4 Z_u^2 + 24Z_+ Z_-^6 Z_u + 128Z_+^4 Z_-^2 Z_u^2 - 128Z_+^2 Z_-^2 Z_u^2 Z_0^2 - 128Z_+^3 Z_u Z_0^4$$

$$a_8' = Z_-^8 + 16Z_+^2 Z_-^4 Z_u^2 + 8Z_+ Z_-^6 Z_u + 128Z_-^4 Z_u^2 Z_0^2 - 32Z_+ Z_-^4 Z_u Z_0^2 - 128Z_+^2 Z_-^2 Z_u^2 Z_0^2 + 125Z_+^2 Z_u^2 Z_0^4$$

$$a_0 = \frac{a_0'}{16Z_-^4 Z_u^2 Z_0^2}, a_2 = \frac{a_2'}{16Z_-^4 Z_u^2 Z_0^2}, a_4 = \frac{a_4'}{16Z_-^4 Z_u^2 Z_0^2}, a_6 = \frac{a_6'}{16Z_-^4 Z_u^2 Z_0^2}, a_8 = \frac{a_8'}{16Z_-^4 Z_u^2 Z_0^2}$$

The transfer function with the use of Chebyshev function is shown in equation (2-14) which is corresponding to this short-circuit stub resonator of UWB filter.

$$H_N(\theta) = \frac{T_N(\cos(\theta)/\cos(\theta_c))}{d \sin^3(\theta)} \quad (2-14)$$

Similar to the last section, $N=5$ has been chosen and (2-14) can be expressed in trigonometric polynomial form in equation (2-15).

$$H_5(\theta) = \frac{1}{d \sin^3(\theta)} \left(\frac{16 \cos^5 \theta}{\cos^5 \theta_c} - 20 \frac{\cos^3 \theta}{\cos^3 \theta_c} + 5 \frac{\cos \theta}{\cos \theta_c} \right) \quad (2-15)$$

After some mathematical derivation in terms of Richard's variable as presented in last section, the transfer function of the Chebyshev function is shown in (2-16) by substituting the equation (2-14) into (2-8).

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$$|S_{21c}|^2 = \frac{(1+S^2)^2 S^6}{b_0 + b_2 S^2 + b_4 S^4 + b_6 S^6 + b_8 S^8 + S^{10}} \quad (2-16)$$

$$b_0 = (r^2 + 25r^2 S_c^2 + 200r^2 S_c^4 + 560r^2 S_c^6 + 640r^2 S_c^8 + 256r^2 S_c^{10}) / d^2$$

$$b_2 = (-20r^2 - 300r^2 S_c^2 - 1080r^2 S_c^4 - 1440r^2 S_c^6 - 640r^2 S_c^8) / d^2$$

$$b_4 = (110r^2 + 630r^2 S_c^2 + 1080r^2 S_c^4 + 560r^2 S_c^6) / d^2$$

$$b_6 = (d^2 - 100r^2 - 300r^2 S_c^2 + 200r^2 S_c^4) / d^2$$

$$b_8 = (25r^2 + 2d^2 + 25r^2 S_c^2) / d^2$$

Figure 2-11. Calculated S_{21} -and S_{11} -parameters of the Chebyshev transfer function for the short-circuit stub resonator UWB filter.

The new Chebyshev transfer function is plotted in Figure 2-11 to show the UWB filter performance with cutoff parameter $\theta_c = 45^\circ$ and ripple value $r = 0.2$ ($L_{Ar} = 15.37$ dB). Comparing with the Figure 2-7, Figure 2-11 has a sharper rejection skirt as the degree of

transmission zeros S increase from 2 to 6 in (2-16), corresponding to the real roots at 0° and 180° . From the transmission line point of view, the high rejection is due to one more pair of transmission zero that is created in the quarter wavelength short-circuit stubs at lower and upper stopbands out of the UWB passband. Nevertheless, since the rejection at cutoff edge becomes sharp, the return loss is slightly deteriorated near the cutoff edge. To implement this Chebyshev function UWB filter by transmission line components, the coefficients of the polynomials in equation (2-13) needs to be correspondingly equal to equation (2-16), e.g. $a_i=b_i$ ($i=0,2,3,4,6$). As a result, the transmission line equation parameters can be solved to achieve same performance as shown in Figure 2-11. Similar to the previous section, three variables Z_{oe} , Z_{oo} and Z_u have to be determined. The first three equations can be utilized to solve these three unknowns so that the passband response of this circuit can be same as in Figure 2-7. By solving the system equations in Matlab, $Z_{oe}=145.4098\ \Omega$, $Z_{oo}=43.7095\ \Omega$ and $Z_u=26.4359\ \Omega$ can be obtained.

Figure 2-12. Optimized layout and dimensions for the EM simulation.

Figure 2-13. The S_{21} - and S_{11} -parameters of circuit and EM simulation

The calculated circuit parameters were then converted to physical dimensions. The optimized layout is shown in Figure 2-12. The calculated circuit parameters were also utilized as the design parameters of the ideal circuit simulation in ADS, the $|S_{11}|$ and $|S_{21}|$ were plotted in Figure 2-13 (solid lines). The passband is almost the same as that in Figure 2-11 with return loss around 15.3 dB. The layout of the UWB filter in Figure 2-12 was optimized by Full-wave simulation in Momentum and its $|S_{11}|$ and $|S_{21}|$ were also plotted in Figure 2-13 (dotted lines). As it can be observed in Figure 2-13, the EM simulation result (dotted lines) is matched with the ideal circuit simulation results (solid lines). As discussed in previous section, there are discrepancies between them because the ohmic loss, radiation loss and frequency dispersion has been considered in EM simulation.

Chapter 3

Lowpass Filter Embedded UWB Filter

3.1 Introduction

In the literature review, the author found that most of the UWB bandpass filters couldn't achieve good upper-stopband performance. In this chapter, several UWB bandpass filters are proposed by using the concept of lowpass filter to suppress the upper-stopband. Two approaches have been used in this chapter: one is to directly cascade the lowpass filter with the existing UWB bandpass filter; another one is to embed a lowpass filter to the MMR of the existing UWB filter. The first approach is straightforward and easy to design, however, the drawback is that it enlarges the circuit size and also increases the insertion loss in the passband. In terms of improved performance and size miniaturization, the second approach is more preferable. In the following sections, the author will discuss both approaches. The author will also show the attractive features of LPF-embedded UWB bandpass filter to improve the upper-stopband and miniaturize the size of UWB bandpass filter and explains and discusses its relevant design procedure.

3.2 Integration of Lowpass Filter and Bandpass Filter

As introduced in the literature review of Chapter 1, a integration lowpass and highpass filter can always constitute of a good upper-stopband with the drawback of large circuit size. Thus the bandpass filter can be cascaded with lowpass filter to achieve good upper-stopband. Figure 3-1(a) depicts the schematics of the initial UWB bandpass filter based on multiple-mode resonator [16], [20], namely, Filter-A. This filter consists of a single resonator in the center section and one microstrip interdigital coupled line placed at each side. As pointed out in [16], this multiple-mode resonator itself is formed with one $\lambda/2$ low-impedance line in the center and two identical $\lambda/4$ high-impedance line sections at the two sides with respect to the ultra-wideband central frequency, i.e., 6.85 GHz. As a result, only the first three resonant modes can be properly placed inside the concerned passband.

Figure 3-1. Schematics of the MMR-based UWB bandpass filters. (a) Filter-A. (b) Filter-B. (c) Filter-C. [62]

Figure 3-1(b) and (c) illustrate the two modified ultra-wideband bandpass filters, namely, Filter-B and Filter-C, respectively. They consist of the initial bandpass filter Filter-A and

two 3rd- or 5th- order lowpass filters in the left- and right-side feeding line sections. High-order lowpass filters have been properly designed in priority with the cut-off frequency slightly higher than 10.6 GHz. The performance of 5th-order lowpass filter is shown in Figure 3-2. Thus, they were supposed to effectively reject the unwanted spurious passbands (>10.6 GHz) happened in the initial bandpass filter. In the design, the two 3rd- or 5th- order lowpass filters were initially characterized using the well-established filter synthesis approach to roughly estimate their overall dimensions under the 3-dB cut-off frequency of about 10.6 GHz. Then, their actual performance and dimensions were numerically derived using the full-wave simulator, i.e., Agilent ADS-Momentum.

Figure 3-2. Predicted S_{21} - and S_{11} -magnitudes of the embedded lowpass filter.

Figure 3-3. Predicted S_{21} -magnitudes of Filter-A, Filter-B and Filter-C. [62]

Figure 3-2 shows S_{11} - and S_{21} -magnitude of 5th order lowpass filter as shown in Figure 3-1(c), which was designed to reject the unwanted passband at upper-stopband. Figure 3-3 indicates the simulated frequency-dependent S_{21} -magnitudes of the three UWB bandpass filters shown in Figure 3-1. First of all, one can find that all of the filters have achieved an almost unchanged passband with the maximum insertion loss around ~ 2.0 dB located near the upper-end of UWB passband. As studied in [16], the realized wide passband was primarily constructed with the reallocation of the first three resonant modes of the MMR inside the passband. As the frequency increases from 3.1 to 10.6 GHz, the electrical thickness of dielectric substrate is gradually increased, thus unfortunately leading to an increment in the radiation-dominated insertion loss in the upper passband-end. The lowest spurious of S_{21} -magnitude for the Filter-A appears around 13.7 GHz as shown in Figure 3-3. This is caused by the 4th-order resonant mode of the constituted multiple-mode resonator [16]. Moreover, in the same case (Filter-A), a spurious wide passband with low insertion loss appears above 16.7 GHz and it is contributed by the 5th- to 7th-order resonant modes of this resonator as well as the 2nd-order coupling peak of a dispersive parallel coupled-line.

As the two identical 3rd-order lowpass filters are simultaneously appended in Filter-B, the S_{21} -magnitude in the spurious passband above 16.7 GHz obtains a significant reduction below -13.0 dB. It is primarily dominated by the high-band rejection of these two lowpass filters. Meanwhile, the upper rejection skirt beyond the passband becomes much sharper since the upper-band rejection of the initial bandpass filter and the two external lowpass filters are cumulative-added. By installing two 5th-order lowpass filters in Filter-C, as shown in Figure 3-2, the S_{21} -magnitude is further suppressed below -25.0 dB over the plotted upper-stopband (11.4 to 20.0 GHz) and the upper rejection skirt around 11.0 GHz is sharpened. In other words, the filters with more-order lowpass structures obtain more deepened and widened upper-stopband.

Figure 3-4. Predicted group delay of Filter-A, Filter-B and Filter-C. [62]

For normal bandpass filters with a few percentages of fractional bandwidth, only their S-parameter magnitudes needs to be investigated. But it is not the case for ultra-wideband filter with about 110% fractional bandwidth. To evaluate such a wideband performance

including phase linearity, the group delay has to be analyzed, which is desirable to be constant within passband. The group delay can be calculated by following equation:

$$\begin{aligned}\Phi_{21} &= \text{Arg}(S_{21}(j\Omega)) \\ \tau_d(\Omega) &= -\frac{d\Phi_{21}(\Omega)}{d\Omega}\end{aligned}\quad (3-1)$$

where the phase response $\Phi_{21}(\Omega)$ is in the unit of radians and frequency variable Ω is in radians per second. Figure 3-4 depicts the predicted group delay of the designed ultra-wideband filters in a frequency ranging from of 1.0 to 13.0 GHz. One can observe from Figure 3-4 that as the order of the external lowpass filters increases, the overall curves of Filter-B and -C gradually rise up due to an additional non-linear contribution caused by the lowpass filters. For a comparative study, the group delay at the center frequency close to 6.85 GHz is found to increase from 0.39, 0.50, to 0.62 ns. On the other hand, the two peaks always appear in the lower and upper ends of the passband in all filters (Filter-A, B, C) and their values are raised in an accelerating manner as the high-order lowpass filters are externally cascaded. In particular, the upper-side peak is observed to always exceed its lower-size counterpart due to the sharper roll-off skirts at upper-stopband.

Figure 3-5. Photograph of Filter-C. [62]

Figure 3-6. Predicted and measured S_{21} -magnitudes of Filter-C. [62]

Figure 3-7. Predicted and measured group delay of Filter-C. [62]

To give an evident support on the above-predicted wide passband performance, ultra-wideband bandpass filters based on Filter-C was fabricated and measured. Figure 3-5 is the photograph of the fabricated Filter-C. Figure 3-6 and 3-7 depict the predicted and measured frequency responses of S_{21} - and S_{11} -magnitudes as well as group delay over the plotted 1.0-to-20.0 and 1.0-to-13.0 GHz ranges respectively. Both sets of results were found to be very reasonable agreement with each other. In particular, our concerned S_{21} -magnitude in measurement was kept below -20 dB over a frequency range from 11.4 to 20 GHz, thus

verifying that a wide and deep upper-stopband had been achieved as expected in theory. Meanwhile, it was found that the measured S_{21} -magnitude achieved -1.0 dB at minimum within the passband and the group delay was about 0.62 ns at center frequency 6.85 GHz. The fractional bandwidth of measured result is slightly smaller than the simulated result with 5.8% difference. This discrepancy is due to the fabrication tolerance, since the interdigital coupled-line required very narrow gap (0.05 mm) to provide tight coupling, which is close to the limit of fabrication tolerance.

3.3 Lowpass-Filter-Embedded MMR UWB Filter with Upper-stopband Improvement

In Section 3.2, the two ultra-wideband bandpass filters (Filter-B and Filter-C) with improved upper-stopband performance were successfully constructed via directly cascading the two lowpass structures with the initial ultra-wideband bandpass filter Filter-A. Nevertheless, this simple type of hybrid ultra-wideband filters definitely increased the overall dimensions to a great extent. This drawback may be overcome by internally embedding a lowpass filter into the multiple-mode resonator in the center so as to construct a compact filter as shown in Figure 3-8, namely Filter-D. In this way, the lowpass-filter-embedded multiple-mode resonator has to be properly characterized to allocate the first few resonant modes within the wide passband while fully removing out the other high-order resonant modes in the upper-stopband (10.6 to 20.0 GHz). Figure 3-8 shows the layout of this filter with marked dimensions. To keep a proper tight coupling degree and a low conductor loss, the slot and strip widths were fixed as 0.05 and 0.10 mm, respectively. The 5th-order lowpass filter, which had designed in Figure 3-2, with the cut-off frequency close

to 10.6 GHz (upper-end of UWB) and utilized to replace the $\lambda/2$ low-impedance line in Filter-A and [16]. Meanwhile, the remaining coupling sections at the two sides were kept unchanged as those in Filter-A. After optimization, the overall length was found to be reduced from 17.2 mm (Filter-A) or 27.4 mm (Filter-C) to 12.9 mm (Filter-D).

Figure 3-8. Schematic of the proposed UWB filter: Filter-D

Figure 3-8 indicates the simulated frequency-dependent S_{21} -magnitudes of the 5th-order lowpass filter shown in Figure 3-2. The results of this lowpass filter show a good lowpass behavior with a low insertion loss in the range lower than 10.0 GHz and a wide stopband with the upper-end exceeding 20.0 GHz. This lowpass filter was then connected with two high-impedance parts at each side, thus making up a composite multiple-mode resonator as shown in Figure 3-8. As the coupling degree was gradually increased via strengthening the length of the interdigital coupled-lines [16], [20], the low-loss ultra-wideband filtering performance was achieved as shown in Figure 3-9. The optimized S_{21} -magnitude curve (solid-line) illustrates that this filter has realized a loss-reduced passband (3.1 to 10.6 GHz) with the maximum insertion loss of 0.6 dB against 2.0 dB in Filter-A to Filter-C. It was because the radiation loss is significantly reduced as the low-impedance section with a wide strip width in Filter-A was replaced by this lowpass structure. In addition to a sharpened

rejection skirt, the upper-stopband was then extended beyond 20.0 GHz with an insertion loss higher than 35.0 dB in a range of 12.5 to 20.0 GHz which was even better than that in Filter-C.

Figure 3-9. Predicted S_{21} - and S_{11} -magnitudes and S_{21} -magnitudes under weak coupling of the Filter-D.

In order to provide a physical insight into the multiple-mode resonance behaviors of this composite multiple-mode resonator, the predicted S_{21} -magnitude (short dotted-line) under the weak coupling case where $L_1=0.1$ mm was also plotted in Figure 3-9. The curve reveals the locations of the first five resonant frequencies over a frequency range between 1.0 to 20.0 GHz, i.e., $f_1=4.2$ GHz, $f_2=6.8$ GHz, $f_3=9.1$ GHz, $f_4=10.6$ GHz and $f_5=18.9$ GHz, respectively. Following the comprehensive discussion in [16], the passband in Figure 3-9 was realized with the use of the first four resonant frequencies ($f_1 \sim f_4$). The 5th resonant frequency (f_5) may intuitively contribute the 1st spurious passband, but its appearance was effectively suppressed by the weakened coupling degree of the interdigital coupled-lines installed at the two sides and the lowpass property of the MMR as discussed in Figure 3-2.

Thus, a very low loss ultra-wideband passband was achieved with a widened and deepened upper-stopband that was extended beyond 20.0 GHz.

Figure 3-10. Photograph of Filter-D.

Figure 3-11. Predicted and Measured S_{21} -magnitudes of Filter-D.

Figure 3-12. Predicted and Measured group delay of Filter-D.

Figure 3-10 shows the photograph of the fabricated Filter-D with the overall length of 12.9 mm against 27.4 mm of the Filter-C denoted in Figure 3-5. Figure 3-11 and Figure 3-12 depict the comparison between the predicted and measured frequency responses of S_{21} - and S_{11} -magnitudes as well as group delay. Good agreement has been achieved again over a frequency ranging from 1.0 to 20.0 GHz. It can be observed that in the UWB passband, the measured results realized a better performance than those of Filter-C in terms of the return/insertion losses and group delay. At the center of passband or 6.85 GHz, the measured S_{21} -magnitudes achieved less than -0.9 dB while the group delay is around 0.23 ns with maximum variation of 0.41 ns, thus showing the lower loss and better phase linearity than those studied in Section 3.2. Furthermore, the measured S_{21} -magnitudes was always kept below -23.5 dB in the concerned upper-stopband covering from 12.0 to 20.0 GHz to make up a widened and deepened upper-stopband beyond the UWB passband. Table I lists the key parameters of aforementioned four UWB bandpass filters. As it can be observed, the Filter-D has the lowest insertion loss, the best upper-stopband and the smallest physical size.

Parameters	Insertion loss within passband (at 6.85 GHz)	Upper-stopband rejection	Physical length
Filter-A	0.49 dB	14 dB (12 GHz -16 GHz)	17.2 mm
Filter-B	0.59 dB	14 dB (11.3 GHz- >20 GHz)	23.6 mm
Filter-C	0.88 dB	20 dB (11.8 GHz – >20 GHz)	27.4 mm
Filter-D	0.41 dB	20 dB (11.5 GHz- >20 GHz)	12.9 mm

Table I. Comparison of key parameters of four proposed UWB bandpass filters

3.4 Miniaturization of LPF-Embedded MMR UWB Bandpass Filter with Upper-stopband Improvement

Figure 3-13. Photograph of miniaturized UWB filter. [42]

Figure 3-14. S_{21} -magnitudes of cell-A, cell-B elements and stubs-loaded MMR. [42]

In this section, the author tried to miniaturize the LPF-embedded MMR UWB filter while keeping the properties of good UWB passband and wide upper-stopband. A modified MMR resonator was formed by transversely attaching three pairs of non-uniform and

folded stubs. Its configuration was then investigated, aiming to allocate the first three resonant frequencies within 3.1-10.6 GHz band while suppressing spurious harmonics at the upper-stopband. Plus, its high-impedance sections in two sides were linked with two feed lines via interdigital coupled-lines, leading to construct an initial UWB filter. Then, two-stage UWB bandpass filter was constructed to sharpen the rejection skirt and deepen the lower and upper-stopband. Finally, two filters were optimized, fabricated and measured to give confirmation on the improved UWB performance.

Figure 3-13 depicts the geometrical schematic of the proposed UWB bandpass filter. This filter consists of the stub-loaded multiple-mode resonator in the center section and two identical interdigital coupled-lines in the left and right sections. Instead of a traditional parallel coupled-line as discussed in section 2.1.2 and 2.1.3, an interdigital coupled line or double parallel-coupled line was used herein to achieve the requested tight coupling degree in the case of a thin substrate thickness, i.e., $h=0.635$ mm [13]. Three pairs of stepped-impedance stubs, namely, cell-A in center and cell-B in two sides, were transversely attached to the middle of the MMR. These stubs were utilized to produce varied wide stopbands above the UWB passband as illustrated in the simulated S_{21} -magnitudes in Figure 3-14.

In the design, the dimensions of one cell-A and two cell-B stubs were properly adjusted to allocate their 3-dB cut-off frequencies near 11.0 GHz. As shown in Figure 3-14, the middle section of the constituted MMR exhibits an excellent lowpass property with sharpened rejection skirt and wide stopband of larger than 20 dB over 13.0-29.0 GHz. As

this MMR was linked with external feed lines via interdigital coupled-lines as shown in Figure 3-14, a desired UWB passband with wide upper-stopband can be expected.

Figure 3-15. S_{21} -magnitudes of UWB bandpass filter in Figure 3-13 with $d=0.6$ and 4.3 mm. [42]

Figure 3-15 plots the simulated S_{21} -magnitude of the filter structure in Figure 3-13 under varied coupled-line lengths, i.e., $d=0.6$ and 4.3 mm, respectively. As discussed in [16], we understood that the first three resonant frequencies, ranging in the 3.1 - 10.6 GHz band, can be effective utilized to make up the desired UWB passband. As the coupled-line length (d) increases, the S_{21} -magnitude curve in the UWB band gradually rises up towards the ideal 0 -dB line. Figure 3-15 shows that an UWB passband is realized under the case of $d=4.3$ mm (solid line) and the bandwidth of the constituted passband is primarily controlled by the resonant frequencies of the first three resonant modes, i.e., f_{m1} , f_{m2} and f_{m3} [16].

In the meantime, the 4th-order mode f_{m4} occurs around 12.7 GHz that is close to the coupling transmission zero of two interdigital coupling sections. Thus, as discussed in [16],

the unwanted passband due to this 4th-order mode can be entirely eliminated as illustrated in Figure 3-15 (solid line). In addition, lowpass behavior of the three paired stubs takes a primary role in highly rejecting the unwanted spurious passbands f_{m5} and f_{m6} at 18.0 and 25.0 GHz, respectively. As a result, an UWB bandpass filter with very wide stopband is achieved in the range from 11.0-26.0 GHz with an insertion loss larger than 20 dB.

Figure 3-16. Photograph of the proposed two-stage UWB filter.

On the other hand, in order to further improve the rejection skirt and meet the FCC defined mask, two-stage UWB bandpass filter was constructed using the MMR structure in Figure 3-13. In the design, two adjacent stubs-loaded-MMRs were coupled via parallel-coupled microstrip lines. The gap spacing and coupled-line length of these coupled-lines were properly adjusted, aiming to achieve a high return loss (>10 dB) within the UWB passband. At the same time, the transmission zero of this coupled-line was moved to effectively reject the 4th-mode-caused harmonic passband.

The final layout of this optimized two-stage UWB filter can be found in Figure 3-16. It has a compact size with the length of 17.01 mm against 27.7 mm [20] under the same two-stage cases. The simulation results, as shown in Figure 3-19, achieved return loss larger than 15 dB in the entire UWB passband. Of course, as the number of MMR was increased,

cumulative transmission loss gradually became significant at higher frequency. In our case, the simulated insertion loss around 10.0 GHz was about 1.7 dB. However, with the use of this two-stage bandpass filter, the FCC defined indoor mask can be well satisfied as shown in Figure 3-19 in the plotted frequency ranging from 1 GHz to 30 GHz.

To provide an evident verification on the above predicted bandpass performances of one- and two-stage UWB filters, two prototype filter circuits were fabricated. Figure 3-17 and 3-19 show the comparison between the predicted and measured frequency responses of S_{21} - and S_{11} -magnitudes for one-stage and two-stage UWB filters respectively. Generally speaking, both sets of the results were found in very reasonable agreement with each other over the plotted range (1.0 to 30.0 GHz). Especially, the measured S_{21} -magnitude consistently became lower than -20 dB in the frequency range higher than 11.0 GHz, thus ensuring that an expected wide and deep upper-stopband has been really realized.

Figure 3-17. Simulated and measured S_{21} - & S_{11} - magnitudes of 1-stage UWB filter (0-30 GHz).

Figure 3-18. Simulated and measured S_{21} - & S_{11} - magnitudes and group delay of 1-stage UWB filter (1-12 GHz).

In order to show the passband performance clearly, the S-parameter results were plotted again in the narrowed frequency ranging from 1-12 GHz together with the group delays. As illustrated in Figure 3-18 and 3-20, again, the predicted and measured results were found in good agreement with each other. The measured insertion loss achieved less than 0.5 dB in one-stage filter and less than 0.8 dB in two-stage filter at the UWB center frequency 6.85 GHz. The group delays, as plotted in Figure 3-18 and 3-20, show a flat frequency response in the passband. It implies good passband linearity with about 0.20 ns and 0.36 ns at 6.85 GHz (UWB's center) in one- and two-stage filters, respectively.

Figure 3-19. Simulated and measured S_{21} - & S_{11} - magnitudes of 2-stage UWB filter (0-30 GHz).

Figure 3-20. Simulated and measured S_{21} - & S_{11} - magnitudes and group delay of 2-stage UWB filter (1-12 GHz).

3.5 Conclusion

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In this section, a class of UWB bandpass filters with improved upper-stopband performance was proposed, designed and implemented. Two identical interdigital coupled lines with enhanced coupling degree were first connected with a half-wavelength multiple-mode resonator (MMR) at two sides to make up an initial wideband filter with the passband that covers the whole UWB band, i.e., 3.1 to 10.6 GHz. By installing two high-order lowpass filters at the left and right sides of this initial filter, the resultant UWB bandpass filter was constructed with effective suppression of unexpected spurious passband that was caused by the 4th- to 7th-order resonant modes in the MMR. After theoretical study was comparatively executed on the three UWB filters, a prototype filter circuit with two 5th-order lowpass filters was finally fabricated and measured to provide an experimental demonstration of its actual UWB passband and upper-stopband performances. In addition, the 5th-order LPF was embedded into the MMR of bandpass filter to replace the uniform half-wavelength resonator. Thus, a reasonably good UWB passband was realized, the measured upper-stopband had achieved an insertion loss of higher than 20 dB and an extremely widened range of 11.4~ 20.0 GHz against 11.3~13.8 GHz in the original UWB filter reported in [16] and [20].

Later on, one- and two-stage LPF-embedded MMR UWB bandpass filters with improved upper-stopband and miniaturized size were proposed, designed and implemented. The simulated and measured results were found to be reasonable matching with each other in the concerned frequency range of 1.0-30.0 GHz. A good UWB passband was realized with measured insertion loss less than 0.5 dB in one-stage and 0.8 dB in two-stage at the UWB's central frequency of 6.85 GHz. The measured upper-stopband achieved an

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insertion loss higher than 20 dB with an extremely widened frequency range of 11.0-26.0 GHz. In addition, the two-stage UWB bandpass filter met the FCC indoor spectrum mask in frequency range of 1.0-30.0 GHz.

Chapter 4

UWB Bandpass Filter with EBG- Embedded MMR

4.1 Introduction

Various electromagnetic bandgap (EBG) structures for microwave circuits have been introduced extensively over the last few years [9], [37], [63], [64]. EBG has been widely used to prevent excitation of higher-order modes of microstrip antennas. It also has been used in filter design to reject higher order passbands, due to its inherent stopband behavior [9], [37], [63], [64]. In this section a class of ultra-wideband microstrip-line bandpass filters with improved out-of-band behaviors is presented by using EBG-embedded multimode resonator (MMR). At first, two EBG structures were constructed by periodically loading uniform or non-uniform open-circuited stubs. Their performances were numerically characterized, aiming at exhibiting the existence of a bandstop in a frequency range of 15.0-to-30.0 GHz. These bandgap structures were then internally embedded in the middle part of an open-ended microstrip line resonator, constituting a modified type of MMR. A UWB

bandpass filter with this MMR was designed. Furthermore, the author extended the EBG-embedded MMR UWB filter to two-stage to dramatically improve the rejection skirts of passband.

4.2 Study of Two Proposed EBG Structures

Figure 4-1. Geometry of two types of three-cell EBG structures loaded by two different open-circuited stubs (Dimensions: $H_1=2.4$, $T_1=1.5$, $T_2=1.15$, $H_2=2.1$ in mm). [64]

Figure 4-2. Equivalent circuit model of three-cell EBG structures. [64]

As well known in theory, there always exist an infinite number of resonant modes for a transmission line resonator. As the first three resonant modes were used to constitute a wide passband, all the remaining high-order modes unfortunately contribute to the emergence of undesired harmonic passbands. To suppress these upper-band harmonics, an upper-

stopband was achieved by externally integrating two identical lowpass filters with the initial UWB filter as discussion in chapter 3. In [37], a wideband filter with extended upper-stopband was constituted by internally installing electromagnetic bandgap (EBG) transmission line into the half-wavelength resonator. This EBG-embedded approach has significantly reduced the overall dimension of the constituted UWB filters.

The objective of this work is to improve the out-of-band behavior of the UWB filters at no extra cost of size enlargement by internally embedding EBG structures into a multimode resonator. To do it, the two microstrip-line EBG structures were comparatively studied in terms of fundamental per-unit-length transmission parameters as discussed in [65]. The numerical results demonstrated their distinctive guided-wave characteristics, i.e., slow-wave at low frequency band and band-gap at high frequency band. These two EBG structures were then internally embedded in the middle part of an open-circuited resonator, aiming to constitute a harmonic-exclusive multimode resonator. As the parallel-coupled lines were introduced with a maximum coupling at central frequency, two ultra-wide bandpass filters based on single- and double-stage resonators were designed, implemented and fabricated. Measured results evidently validated predicted results, exhibiting improved out-of-band upper-stopband with an insertion loss (>15 dB in 11.8-30.0-GHz range), sharpened roll-off rejection skirts and miniaturized overall length.

Figure 4-1 depicts the two types of electromagnetic bandgap (EBG) structures to be considered. The unit cell of EBG-I structure was constituted by transversely attaching a pair of uniform open-circuited stubs in the middle of a microstrip-line section of the length (T_1). By making effective use of the slow-wave property of EBG-I structure, the constituted non-

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uniform resonator was much smaller or shorter in length than the uniform resonator. The length of L_p along the EBG section was selected to set up its bandstop to be located higher than the UWB passband, aiming at suppressing harmonic passbands as exhibited in [16]. These stubs can be considered as a simple shunt capacitance at extremely low frequencies, but need to be modified as a generalized shunt reactance at relatively high frequencies with the capacitance of $2C$ and inductance of $L/2$ as shown in Figure 4-2. More strictly speaking, this equivalent shunt reactance can still be valid at not so high frequencies, in which the stub lengths are electrically short and the substrate thickness is much smaller than its operating wavelength.

Figure 4-3. Extracted reactance $img(Z_{12})$ from fullwave EM simulation for the two type of EBG structures. [64]

Regarding the extraction of these lumped elements, the shunt reactance $img(Z_{12})$ of these stubs was calculated via Agilent full-wave ADS software and Figure 4-3 plots its values as a function of frequency. To solve these two unknown parameters in this model, i.e., $2C$ and $L/2$, the two values of such a reactance at two low frequencies, e.g., 3.0 and 5.0

GHz, need to be selected to establish a couple of equations with reference to the extraction approach used in extracting equivalent circuit model for a spiral inductor in [66] and that for a microstrip open-ended discontinuity.

For the structure of EBG-I, a couple of equations were formed as follows

$$\text{img}(Z_{12})\Big|_{f_1=3\text{GHz}} = j\varpi_1 L/2 + 1/(j\varpi_1 2C) = -j127 \quad (4-1a)$$

$$\text{img}(Z_{12})\Big|_{f_2=5\text{GHz}} = j\varpi_2 L/2 + 1/(j\varpi_2 2C) = -j76 \quad (4-1b)$$

Similarly, for the structure of EBG-II, these two equations became

$$\text{img}(Z_{12})\Big|_{f_1=3\text{GHz}} = j\varpi_1 L/2 + 1/(j\varpi_1 2C) = -j130 \quad (4-2a)$$

$$\text{img}(Z_{12})\Big|_{f_2=5\text{GHz}} = j\varpi_2 L/2 + 1/(j\varpi_2 2C) = -j75 \quad (4-2b)$$

Figure 4-4. Simulated S_{21} - and S_{11} -magnitudes in TX model and EM. (TX: circuit model; EM: electromagnetic simulation). [64]

Figure 4-4 shows two groups of simulated results based on the calculated circuit model (TX) in Table I and electromagnetic simulator (EM). They were very reasonably matched with each other in the range below 18.0 GHz. Nevertheless, there was a large discrepancy at frequencies higher than 18.0 GHz as can be seen from Figure 4-4. Based on basic electromagnetic theory and our previous works in parametric extraction of planar circuits such as [67] using the full-wave method of moments, it can be explained as follows. On one hand, the shunt stubs with uniform and stepped-impedance configurations, as shown in Figure 4-1, can only be reasonably expressed as an equivalent simple lumped-element LC network at low frequencies. As the frequency increases to a high value such as 18.0 GHz, these stubs become electrically long so as to bring out frequency dispersive characteristics. For an instance, at 18.0 GHz, the guided-wavelength of the uniform stub in Figure 4-3 is equal to about 5.4 mm. Therefore, this stub with $H_1=2.4$ mm given in Table I is no longer much shorter than the wavelength at and above 18.0 GHz. On the other hand, as the frequency increases to the extent that the wavelength is comparable to the thickness of a substrate, a number of unexpected EM effects emerge, inclusive of frequency dispersion, radiation loss and high-order surface-wave modes.

EBG-I				
W_1	T_1	H_1	$L/2=L_1/2$	$2C=2C_1$
0.7 mm	1.5 mm	2.4 mm	20 pH	0.42 pF
EBG-II				
W_2	T_2	H_2	$L/2=L_2/2$	$2C=2C_2$
0.7 mm	1.15 mm	2.1 mm	110 pH	0.4 pF

Table I. Dimensions and extracted parameters of two EBG structures in Figure 4-4 under identical cut-off frequency of 13 GHz. [64]

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Following the previous studies in [37] and [65], the EBG-II structure was constituted by introducing a pair of stepped-impedance open-circuited stubs to sharpen the roll-off skirt near the cut-off frequency. Calculated S_{21} - and S_{11} -magnitudes of this EBG-II were also plotted in Figure 4-4. It was seen herein that the first transmission zero, caused by resonant frequency in EBG-II, was lowered to the cut-off frequency of 13.0-14.0 GHz. It actually helped us achieve a sharper and deeper rejection slope. In addition, the overall size was found to be reduced by 33% if the EBG-II was used. Table I tabulates the extracted circuit parameters of these two EBG structures. It is shown that the shunt inductance L_1 in EBG-I is much smaller than its counterpart L_2 in EBG-II. This increased inductance in EBG-II results in lowering the transmission zero and speeding up the roll-off rate of S_{21} -magnitude.

To further investigate the transmission behaviors of these two EBG structures, the guided-wave propagation constants ($\gamma=j\beta+\alpha$) were numerically extracted from the EM simulation of a two-port EBG structure. Following our previous work in [65], the propagation constant can be, in general, expressed in terms of two ABCD-matrix elements, A_p and D_p in equation (4-3).

$$\cosh(\gamma L_p) = \frac{A_p + D_p}{2} \quad (4-3)$$

where L_p is the overall length of a N-cell EBG line and $\gamma=j\beta$ is valid for the lossless case. Figure 4-5 depicts the extracted normalized phase constant (β/k_0) of EBG-I and EBG-II structures with virtue of the transmission line model and electromagnetic (EM) simulation, respectively, where k_0 is the phase constant in free space. They are found to be matched with each other except for the frequencies higher than 24.0 GHz. The discrepancy is caused

by the ignorance of frequency-dispersive effects in the circuit model. As shown in Figure 4-5, β/k_0 in EBG-II is much higher than that in EBG-I at low frequencies (<15.0 GHz), implying that the slow-wave factor of EBG-II is larger than that of EBG-I.

Figure 4-5. Extracted normalized imaginary part propagation constant of EBG-I and EBG-II in ideal TX model and EM simulation with fixed periodic $T_1=1.50\text{mm}$ and $T_2=1.15\text{mm}$. [64]

On the other hand, as the frequency exceeds 15.0 GHz, the attenuation constant α/k_0 (unit cell of $N=1$) appears above zero as indicated in the shaded region of Figure 4-6. This shaded region in Figure 4-5 and 4-6 represents a frequency band with nonzero α/k_0 , that is referred to as the so-called bandgap or bandstop [65]. Again, the EM-simulated results were reasonably matched with those based on the ideal transmission line models for both EBG-I and EBG-II structures. Using the calculated lumped-element parameters as listed in Table I, the resonant frequency of transverse stepped-impedance stubs in EBG-I is found to be above 30.0 GHz. It can help us understand that α/k_0 has not yet received a significant influence from this stub resonance and thus becomes smoothly varied in the stopband.

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However, this resonant frequency in EBG-II appeared around 26.0 GHz as the inductance ($L_2/2$) was increased, thus creating an attenuation pole in the stopband. This pole in EM model is much smaller than that in TX model due to the fact that the radiation, material and metallic losses as well as complicated coupling effects were all taken into account in the EM model. This high attenuation in the bandstop is used later on to suppress the higher resonant modes existed in the UWB filter discussed in our previous works in section 3 and [16], [37] and [64].

Figure 4-6. Extracted normalized real part propagation constant and $|S_{21}|$ in Bloch-wave impedance of EBG-I and EBG-II in ideal TX model and EM simulation with fixed periodic $T_1=1.50\text{mm}$ and $T_2=1.15\text{mm}$. [64]

Now, let's consider that the guided-wave propagates across a N-cell EBG structure with perfect impedance matching at its two truncated terminals that can be realized by setting the two external feed-line characteristic impedances as its Bloch-wave impedance.

$$S_{21} = e^{-\gamma NT} \quad (4-4)$$

where T is the period of the EBG unit cell. Using equation (4-4), $|S_{21}|$ can be simply calculated as a function of frequency and it was plotted in Figure 4-6 for the EBG-II structure with $N=1, 2$ and 3 . As the number of unit cell N increased, the attenuation in the relevant stopband became higher and higher. Nevertheless, the attenuation pole in α -curves was completely consistent with that in $|S_{21}|$ -curves regardless of varied EBG unit cell (N). The general design procedure of the EBG-I structure was primarily classified into two steps: (a) L_p (length of EBG section) is initially selected higher than the concerned UWB band, e.g., ~ 11.0 GHz; (b) Shunt stub lengths are all selected as about one quarter-wavelength at the 4th- or 5th-order resonant frequencies discussed in [16] and [37].

4.3 Study EBG-Embedded UWB Bandpass Filter

Figure 4-7. Geometries and dimensions of the proposed UWB bandpass filters EBG-I-embedded MMR ($d_1=4.3$ mm). [4]

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Figure 4-8. Geometries and dimensions of the proposed UWB bandpass filters EBG-II-embedded MMR ($d_1=4.25$ mm). [4]

In this section, the above-discussed EBG structures with the unit cell of $N=3$ were internally embedded into the middle of the multimode resonator [16] to replace the uniform low-impedance section. The two filters consist of two distinctive parts, i.e., EBG-embedded multimode resonator in center and two identical $\lambda_g/4$ microstrip interdigital coupled lines at two sides. It needs to be emphasized that all the traditional synthesis design approaches or methodologies are established under the assumption of a narrow passband and a single-mode operation. There does not have any reported circuit-model based synthesis approach available for efficient optimization of such a wideband filter with the 110% bandwidth. Our optimization is carried out in the ADS software following the three main steps: (a) EBG structure with finite cells was studied to produce a wide bandstop above the UWB passband; (b) EBG-embedded multimode resonator were constituted with the three resonant modes in the UWB band; (c) parallel-coupled lines were modeled towards miniaturizing in-band insertion loss and canceling 4th-order resonant mode. Figure 4-7 and 4-8 illustrate the final layouts and dimensions of these two EBG- embedded UWB bandpass filters.

Figure 4-9. Simulated S_{21} -magnitudes of the two proposed UWB bandpass filters in Figure 4-7 and 4-8 under initial weak coupling ($d_1 = d_2 = 0.1$ mm). [64]

Figure 4-10. Simulated S_{21} -magnitudes of the proposed EBG-II UWB bandpass filters in Figure 4-8 under varied coupling lengths of $d_2 = 0.2, 0.6$ and 1.0 mm. [64]

Figure 4-9 plots the two sets of simulated S_{21} -magnitudes of the two filters in Figure 4-7 and 4-8 under weak coupling condition (coupling length: $d_1 = d_2 = 0.1$ mm). It was shown that all the resonant modes of the constituted multimode resonators appeared almost at the same frequencies. The first three resonant frequencies, i.e., f_{m1} , f_{m2} and f_{m3} , were found in the concerned UWB passband. As the coupled-line length was stretched ($d_1 = 4.30$ mm and

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$d_2 = 4.25$ mm), an ultra-wide passband, covering the 3.1-10.6 GHz UWB band, could be constituted as shown in Figure 4-11. It can be imaged from Figure 4-9 that the lower and higher cut-off frequencies of the dominant passband are actually determined by the 1st- and 3rd-order resonant modes. Whereas, the 4th-order mode at f_{m4} is suppressed by coupling transmission zero of two interdigital coupling sections around 13-14 GHz as shown in Figure 4-11 [37].

Figure 4-11. Simulated S_{21} - and S_{11} -magnitudes of the two proposed UWB bandpass filters in Figure 4-7 and 4-8 under optimized tight coupling ($d_1 = 4.3$ mm, $d_2 = 4.25$ mm).[64]

In order to demonstrate how the attenuation at f_{m5} was highly raised, Figure 4-10 was plotted for varied coupling lengths (d_2). As d_2 increases from 0.2, 0.6 to 1.0 mm, $|S_{21}|$ in the first four resonant frequencies rises up toward the 0-dB line due to enhanced coupling degree. Whereas, $|S_{21}|$ at the 5th and 6th resonant frequencies receives a progressive attenuation since the coupling degree in this frequency range becomes weaker and EBG phenomenon take effects. In Figure 4-11, the embedded EBG structures shows that the S_{21} -magnitude attenuates in a wide frequency range above 15.0 GHz, including the two peaks

at $f_{m5}=19.0$ GHz and $f_{m6}=29.0$ GHz, respectively. As a result, an extremely wide upper-stopband covering 11.8 to 30.0 GHz was realized. Looking closely at Figure 4-11, the passband is quite similar to both of the single-stage UWB filters with a high return loss (>16 dB). Thus, it can be concluded that the EBG-II embedded UWB filter is superior to its EBG-I counterpart since it possesses a deep upper-stopband and miniaturized size.

4.4 Predicted and Experimental Results

Figure 4-12. Photograph of the fabricated single-stage UWB microstrip-line bandpass filters by using the EBG-II embedded resonator.[64]

Figure 4-13. Photograph of the fabricated double-stage UWB microstrip-line bandpass filters by using the EBG-II embedded resonator.[64]

In this section, multi-stage UWB microstrip-line bandpass filters were designed, implemented and fabricated. In the design, two adjacent resonators were capacitively coupled via parallel-coupled microstrip lines. The transverse gap spacing and longitudinal

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coupling length of these coupled lines were properly adjusted, aiming to achieve a high return loss (>15.0 dB) in the UWB passband while eliminating the 4th-mode-caused harmonic. The photographs of the two optimized filters are illustrated in Figure 4-12 and 4-13. The overall length of the single-stage UWB filter is about 11.2 mm. It is much smaller than that of the initial UWB filters in [16] due to slow-wave behavior of the embedded EBG structure. For the reader's reference, the overall length of this double-stage filter is about 18.0 mm.

Figure 4-14. Predicted and measured S_{21} - and S_{11} -magnitudes frequency responses of the single-stage UWB bandpass filters.[64]

Figure 4-15. Predicted and measured group delay frequency responses of the single-stage UWB bandpass filters.[64]

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Figure 4-14 and 4-15 are the simulated and predicted frequency responses of S_{21} - and S_{11} -magnitudes as well as group delays of the single-stage UWB filter in Figure 4-12. We can observe from Fig 4-14 that the dominant ultra-wide passband of 3.1-10.6 GHz was satisfactorily achieved with the return loss higher than 11.0 dB. The insertion loss in the upper-stopband of 11.8-30.0 GHz was larger than 16.0 dB. These results proved that the proposed EBG-II embedded resonator could be utilized to make up a modified class of ultra-wideband bandpass filters with extended and deepened upper-stopband. Figure 4-15 shows the group delay in the UWB passband which is plotted from 1.0 GHz to 12.0 GHz. Its shape seems very flat as the frequency varies in the middle range, implying a good linearity in the passband. Nevertheless, it becomes tremendously varied as the frequency moves to the lower and upper cut-off frequencies as reported in [16] and [37]. As compared with the UWB bandpass filter designed in Figure 3-10, although both of the UWB bandpass filters have good passband and wide upper-stopband performance, EBG-II UWB bandpass filter has a much smaller size. The circuit area is 23.52 mm² for the EBG-II UWB bandpass filter against 41.28 mm² for the UWB bandpass filter in Figure 3-10, which has 43% size reduction. The trade-off is that the EBG-II UWB bandpass filter has a more complicate layout structure, whereas the UWB bandpass filter in Figure 3-10 is a traditional 5th-order step impedance lowpass filter.

Figure 4-16 and 4-17 show the predicted and measured frequency responses of S_{21} - and S_{11} -magnitudes as well as group delay of double-stage UWB bandpass filter as shown in Figure 4-13. They were found to be in very reasonably good agreement. The measured insertion and return losses at the UWB center were lower than 1.0 dB and higher than 10.0

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dB, respectively. The measured insertion loss of the upper-stopband was higher than 15.0 dB in the 11.0-30.0 GHz frequency range. Figure 4-16 also shows that as the stage number is increased from one to two, the out-of-band rejection skirts in both lower and upper stopbands become sharpened quickly, thus moving a step towards the FCC-defined UWB mask [16]. Comparing the result in Figure 4-15 and 4-17, the group delay rises as the stage number increases because the overall length is stretched in two-stage case. In particular, the group-delay variation at these frequency edges becomes larger as two resonators are employed. It is predominantly caused by an inevitable tradeoff between the rejection skirt and group delay that exists in any microwave filter block [9], [16], [37], [64] etc. Moreover, it can be expected that as the number of stages increases, the group delay values rise up due to the increase length.

Figure 4-16. Predicted and measured S_{21} - and S_{11} -magnitudes frequency responses of the double-stage UWB bandpass filters. [64]

Figure 4-17. Predicted and measured group delay frequency responses of the double-stage UWB bandpass filters. [64]

Nevertheless, the author could still find some slight discrepancies between the simulated and measured results displayed in Figure 4-14 and Figure 4-16. In fact, it was not easy to realize the tightly coupled-lines involved in the UWB filter design. The coupling gap is usually ranged between 0.025 and 0.075 mm, and it is too small to be accurately fabricated using the etching process since the minimum size that can be well guaranteed for available commercial etching process is about 0.10 mm or 4 mil. The conductor thickness of ~ 0.02 mm is not negligibly small in the case of narrow gap and strip widths, i.e., 0.05 mm and 0.10 mm, affecting the measured performances of the fabricated filters to a certain degree. In addition, it is really challenging and also difficult to carry out a calibration procedure in such a wideband frequency range (1.0-to-30.0 GHz). In our case, a simple two-full-port calibration procedure was executed over the wide frequency range in terms of three available calibration standards, i.e., short, open and through. It is worthy to note, this two-stage EBG-II UWB bandpass filter has a better upper-stopband than the filter introduced in Figure 3-16. As it can be observed, the upper-stopband at 28-30 GHz in Figure 3-17 is deteriorated and the attenuation is reduced to less than 10 dB. However, the

trade-off of wider upper-stopband in Fig. 4-16 is that the physical area of the filter is slightly increased from 37.422 mm^2 to 37.8 mm^2 or 1%.

4.5 Conclusion

In this work, two types of electromagnetic bandgap (EBG) structures, namely, EBG-I and EBG-II, were studied to show their distinctive slow-wave and bandstop behaviors and to make up a modified class of EBG-embedded multimode resonator. Two single-stages UWB microstrip-line bandpass filters with good out-of-band performances were theoretically investigated and designed. Afterwards, a double-stage UWB bandpass filter using the EBG-II was implemented by cascading the two EBG-embedded multimode resonators via parallel-coupled microstrip lines. In the design, these multimode resonators were characterized to allocate their first three resonant modes within the concerned 3.1-10.6 GHz UWB passband. Then, their 4th-order resonant mode was properly placed at the coupling zero of two external coupled-lines. The measured results of the fabricated single- and double-stage UWB filters had very reasonably validated the predicted results in terms of insertion/return losses over a frequency ranging from 1.0 to 30.0 GHz and group delays ranging from 1.0 to 12.0 GHz. It was demonstrated in simulation and experiment that a wide upper-stopband was achieved with the insertion loss higher than 15.0 dB, whereas the out-of-band rejection skirts in both the lower and upper sides became significantly sharpened as the two resonators were used. The group delay was very flat and varied in a range from 0.2 and 0.35 ns in the UWB passband.

Chapter 5

UWB Filter with Short-Circuit Stub Embedded MMR

5.1 Introduction

In the previous two chapters, two categories of UWB filters were discussed; the lowpass filter embedded MMR in chapter 3 and EBG-embedded MMR in chapter 4. Both aforementioned UWB filters have significantly improved the upper-stopband. Nevertheless, their lower-stopband and rejection skirt were not good enough in the case of one-stage MMR filter. In chapter 3 and 4, two-stage MMR UWB filters were designed. They improved the rejection skirt and out-of-band significantly to meet the FCC requirement. The drawback of this two-stage MMR was that the size and the group delay was dramatically increased. In this chapter, the author proposes a short-circuit stub embedded MMR to improve the rejection skirt of the UWB bandpass filter with single MMR.

5.2 Triple-Mode UWB Filter

UWB Filter with Short-Circuit Stub Embedded MMR

The triple-mode UWB bandpass filter was originally reported in [16] and many of the triple-mode UWB filter were also developed in Chapter 3 and 4. As part of the design of a quadruple-mode UWB filter, an initial triple-mode UWB bandpass filter was firstly constituted and characterized to exhibit good harmonic suppression. As a starting part of this work, a modified triple-mode UWB bandpass filter is at first discussed. Figure 5-1 and 5-2 denote its schematic and equivalent transmission line model, respectively. Herein, the two interdigital coupled-lines were formed to provide sufficiently strong coupling degree in the desired UWB band. These two coupled-line sections were then directly tap-connected to the two sides of a uniform half-wavelength resonator. As discussed in [68], a transmission zero could be generated by the tapping technique. Furthermore, its position could be controlled via the tapping position, i.e., length ratio (u/v). Figure 5-3 shows the mode chart for normalized resonant frequencies of the first eight modes and those of a few zeros of the UWB filter in Figure 5-1 with respect to the ratio of u/v . Following the studies in [16], our target herein is to make use of the first three resonant frequencies, f_1 , f_2 and f_3 , to constitute the desired UWB passband while cancelling the 4th one, f_4 , using the 2nd zero of the coupled line, i.e., f_{z2}/f_1 .

Figure 5-1. The schematic of the proposed Type-A triple-mode UWB bandpass filter on microstrip line topology. [39]

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Figure 5-2. The equivalent transmission line model of the proposed triple-mode UWB bandpass filter on microstrip line topology. [39]

Figure 5-3. Mode chart that shows normalized resonant-mode frequencies and transmission zeros versus u/v ratio. [39], [43]

At $u/v=0$, the MMR in Figure 5-1 is almost the same as the original one in [16] in terms of the resonant-mode distribution. The first three modes with f_1 , f_2 and f_3 constitute the desired passband and those with f_5 , f_6 and f_7 form the unwanted harmonic passband. As the value of u/v increases, f_5 , f_6 and f_7 tend to move closer to each other, especially in the range from $u/v = 0.21$ to 0.25 . Meanwhile, the 3rd zero with the normalized frequency, f_{z3}/f_1 , can be excited by adjusting the stub length, u . This gives us a hint to suppress all the resonant modes with f_5 , f_6 and f_7 by properly choosing u/v near the vertical a - a' plane as shown in Figure 5-3. On the other hand, the 8th resonant mode with f_8 can be suppressed by the 3rd zero of the coupled line, i.e., f_{z4}/f_1 , as shown in Figure 5-3. As a result, the triple-mode

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UWB bandpass filter can be worked out in terms of its first three resonant modes. By canceling its 4th to 8th resonant modes, its upper stopband can be significantly extended. In the design, $u/v=0.23$ was found in the a-a' plane. For a fixed length of the vertical part in Figure 5-1, $u=1.1$ mm and $v=4.8$ mm could be obtained.

Figure 5-4. Simulated S_{21} - and S_{11} -magnitudes of the triple-mode UWB bandpass filter with $d=4.25$ and 1.5 mm respectively. [43]

Figure 5-4 shows the simulated S_{21} - and S_{11} -magnitudes of the triple-mode UWB bandpass filter as explained in Figure 5-3 under the selected coupling length of $d=4.25$ (tight coupling case) and $d=1.50$ mm (weak coupling case), respectively. In weak coupling case as discussed in Figure 5-3 and also in [16], the first three resonant modes, ranging in the 3.1-10.6 GHz band, can work together to make up the desired UWB passband. In addition, due to existence of the 3rd zero f_{z3}/f_1 , the f_5, f_6 and f_7 are highly suppressed in both tight and weak coupling cases. From the results in Figure 5-4, it can be figured out that as the coupling length (d) increases, the S_{21} -magnitude curve in the UWB band can gradually rise up towards the ideal 0-dB line. In the case of $d=4.25$ mm (solid line in Figure 5-4), a

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desired UWB passband is satisfactorily realized with the bandwidth dominated by the first three resonant frequencies, i.e. f_1 , f_2 and f_3 . Furthermore, the 2nd to 4th transmission zeros, i.e. f_{z2}/f_1 , f_{z3}/f_1 and f_{z4}/f_1 , have suppressed the 4th to 8th resonant modes. Thus, a wide upper-stopband can be achieved from the higher cut-off frequency to 30 GHz with the rejection larger than 15.0 dB.

Figure 5-5. Geometries of the Type-B triple-mode UWB microstrip-line bandpass filter. [39]

Figure 5-6. Geometries of the Type-C triple-mode UWB microstrip-line bandpass filter. [39]

For the purpose of miniaturization of this triple-mode UWB bandpass filter, Type-A filter is then symmetrically bent to form the Type-B filter as shown in Figure 5-5. Since all the dimensions are kept to be unchanged except the two right-angle corners, the S_{21} -magnitude performance should be very similar to that of the Type-A in Figure 5-4. Nevertheless, there are resonance-caused peaks at around 17.5 and 24.0 GHz that harmfully degrade the upper-stopband performance. To solve this problem, two open-circuited stubs were introduced at the tap-connected position of the Type-B MMR. Thus, a tap-connected

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and stub-loaded MMR, namely Type-C, was formed. Its dimensions were optimized with the target of good in-band and out-of-band performances via ADS software and their values were given in Figure 5-6.

Figure 5-7. Simulated S_{21} - and S_{11} -magnitudes for Type-B ($d=4.25$ mm) and Type-C ($d=4.25$ and 1.5 mm). [39]

Figure 5-7 shows the simulated S_{21} - and S_{11} -magnitudes of the triple-mode UWB bandpass filter under the selected coupling length of $d=4.25$ and $d=1.00$ mm respectively. As discussed in Figure 5-3, the first three resonant modes, ranging in the UWB passband, can work together to make up the desired UWB passband. From the result in Figure 5-7, we can figure out that as the coupling length (d) increases, the S_{21} -magnitude curve in the UWB band can gradually rise up towards the ideal 0-dB line. In the case of $d=4.25$ mm (solid line in Figure 5-7), a desired UWB passband is satisfactorily realized with the bandwidth dominated by the first three resonant frequencies, i.e., f_1 , f_2 and f_3 . Furthermore, the 2nd to 4th transmission zeros, i.e., f_{z2}/f_1 , f_{z3}/f_1 and f_{z4}/f_1 , have fully suppressed the 4th to 8th resonant modes. Thus, a wide upper-stopband can be achieved in the range from the

higher cut-off frequency to 30 GHz with the rejection larger than 20 dB.

5.3 Experimental Result of Triple-Mode UWB Filter

To provide an evident verification on the above-predicted UWB passband and upper-stopband performance, a prototype filter circuit was fabricated and its photo is shown in Figure 5-8. Figure 5-9 shows the comparison between the predicted and measured frequency responses of S_{21} - and S_{11} -magnitudes as well as group delay. Generally speaking, both sets of the results were found to be in very reasonable agreement with each other over the plotted frequency range (1.0 to 30.0 GHz). Especially, the measured S_{21} -magnitude consistently became lower than -17 dB in the frequency range from 12 GHz to 30 GHz, thus ensuring that an expected triple-mode UWB bandpass filter with wide and deep upper-stopband has been really realized. The measured 3-dB cut-off frequencies were 3.4 and 10.3 GHz, thus achieving the fractional bandwidth of about 104%. In addition, within the realized UWB passband, the measured S_{21} -magnitude achieved less than 1.2 dB and the group delay was about 0.22 ns at 6.85 GHz (center frequency of UWB passband).

Figure 5-8. The top-view photograph of Type-C triple-mode UWB bandpass filter. [39]

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Figure 5-9. predicted and measured S_{21} - and S_{11} -magnitudes frequency responses as well as group delay. [39]

A miniaturized triple-mode UWB bandpass filter with extremely wide upper-stopband was designed. By studying different configurations of MMR, proper tap-connection of short open-circuited stubs to a half-wavelength resonator and coupled-line feeding of the constituted MMR have exhibited an attractive capacity in effectively suppressing the higher order resonant modes above the desired UWB passband. This MMR was further miniaturized in terms of bent or curved configuration with good in-band and out-of-band performances. To verify the predicted results, a compact triple-mode UWB bandpass filter was fabricated and tested. Measured results showed us that an extremely wide upper-stopband up to 30 GHz was really achieved in experiment. In the next section, this tap-connected circuit is extended to sharpen its rejection skirts of UWB bandpass filter, whereas remaining a wide upper-stopband.

5.4 Quadruple-Mode UWB Filter

Figure 5-10. The schematic of the proposed quadruple-mode UWB bandpass filter on microstrip line topology. [43]

Figure 5-11. The equivalent circuit model of the proposed quadruple-mode UWB bandpass filter on microstrip line topology. [43]

Figure 5-10 and 5-11 illustrates the schematic and the equivalent transmission line model of the proposed quadruple-mode UWB bandpass filter respectively. The two identical short-circuited stubs with an approximate quarter-wavelength at the center of the desired passband were simultaneously placed at the tapped points of the triple-mode resonator as shown in Figure 5-1. These short-circuited stubs have been successfully applied to the design of an alternative class of UWB bandpass filters as reported in [32]. They were utilized herein to push the 4th resonant mode into the desired passband while sharpening the rejecting skirt of the passband.

Figure 5-12. Simulated S_{21} - and S_{11} -magnitudes for $d=4.25$ and 1.5 mm. [4]

Figure 5-12 shows the simulated S_{21} - and S_{11} -magnitudes of this quadruple-mode UWB bandpass filter under the coupling lengths of $d=4.25$ and 1.50 mm respectively. As demonstrated in Figure 5-12, the 4th-resonant mode is really moved down to its first three counterparts (dotted-line) and it works with them together to make a UWB passband (solid-line). As compared to the results in Figure 5-4 and 5-7, this quadruple-mode UWB filter has enlarged attenuation skirts near the lower and higher cut-off frequencies. In addition, wide upper-stopband performance was still remained.

5.5 Experimental Result of Quadruple-Mode UWB Filter

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Figure 5-13. Top-view photograph of the proposed quadruple-mode UWB bandpass filter. [43]

To provide an evident verification on the predicted passband and upper-stopband performances, a quadruple-mode UWB filter was fabricated and its photograph is shown in Figure 5-13. Figure 5-14 and 5-15 show the comparison between the predicted and measured frequency responses of S_{21} - and S_{11} -magnitudes as well as the group delay. They were found to be in good agreement with each other, thereby the actual realization of the preferred UWB passband and extended upper-stopband was well confirmed in the experiment. Looking at Figure 5-14 closely, the measured upper-stopband is now dramatically extended up to 30 GHz with an insertion loss larger than 15 dB. The measured insertion loss is found less than 1.1 dB within the desired UWB passband. The 3-dB passband covers the range of 2.8-11.0 GHz and it has a fractional bandwidth of 119%. Moreover, the rejection skirt near the cut-off frequencies are sharpened to a great extent against the one-stage MMR-based UWB filters in [1] and those discussed in chapter 3 and 4. In addition, the group delay varies between 0.19-0.52 ns within the UWB passband (e.g. 3.1-10.6 GHz), showing a good linearity.

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Figure 5-14. Predicted and measured S_{21} - and S_{11} -magnitudes frequency responses of the proposed quadruple-mode UWB bandpass filter. [43]

Figure 5-15. Predicted and measured Group delay frequency responses of the proposed quadruple-mode UWB bandpass filter. [43]

5.6 Conclusion

In this work, a triple-mode UWB bandpass filter was first studied and designed with miniaturized size. By tap-connecting a half-wavelength resonator and feeding this MMR via interdigital coupled-lines, its upper-stopband was extended to 30 GHz. Then, by introducing two short-circuited stubs with one quarter-wavelength to this initial MMR, a quadruple-mode MMR was constituted. Based on this modified MMR, a UWB bandpass

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filter was designed to exhibit its attractive sharper rejection skirts and extended upper-stopband without cascading MMRs as discussed in Chapters 3 and 4. A filter prototype was fabricated to demonstrate the predicted UWB performances in the experiment.

Chapter 6

Notch-Band Implementation

6.1 Introduction

In the literature review in the chapter 1, most of UWB bandpass filters have good in-band performance and are suitable for practical implementation. Nevertheless, existing radio signals such as wireless local-area network (WLAN) may have interference with each other in the UWB radio system within the defined UWB frequency spectrum. Thus, a narrow-band notched UWB bandpass filter needs to be developed [33], [35], [69] and [42]. In the following, two approaches of notch-band filter are introduced. The first one is asymmetric coupled arms and the second one is asymmetric shunt stubs. The second approach not only introduces a controllable notch-band, but also sharpens the rejection skirts. After their electrical performances were optimally designed via Agilent ADS Momentum, they were then formed on the RT/Duroid 6010 substrate with a dielectric constant of 10.8, loss tangent 0.0023 and a thickness of 0.635 mm.

6.2 Notch-band implementation using asymmetric coupled-line

Figure 6-1. Schematic of the proposed band notched UWB bandpass filter (all in mm). [42]

In this section, the interdigital coupled lines were re-arranged by lengthening and folding one of two coupling arms associated with the feed lines, so as to create a notch-band at 5.6 GHz within the UWB passband. The UWB filter in section 3.4 was utilized as an example of this UWB bandpass filter with notch-band. As shown in Figure 6-1, two coupling arms now have dissimilar electrical lengths. Thus, the two signal paths along them become out-of-phase and equal-magnitude in a certain frequency in which transmission signal is cancelled, thus generating a notch-band. This approach is applicable to all UWB filter with interdigital coupling section. To demonstrate this performance, a single interdigital coupled-line with different lengths of the folded arms was investigated and its relevant simulated S_{21} -magnitudes were plotted in Figure 6-2. As the length of folded arm varies from 1.60, 1.40 to 1.20 mm, the transmission zero has a gradual increment from 5.25, 5.40 to 5.60 GHz. With this result, it can be figured out that as this length in Figure 6-1 is set up as $h=1.60$ mm (solid line), a notch-band can be introduced at 5.60 GHz within the UWB passband. In optimization design of such a notch-band filter, the author optimized the spacing and length of a standard interdigital coupled-line for minimizing the in-band insertion loss of a UWB bandpass filter in Figure 3-13. Then the stretched length ($2h$) of a

folded arm was adjusted for allocating the transmission zero to the notch band at 5.6 GHz.

Figure 6-2. S_{21} -magnitudes of extracted interdigital coupling section with varied h values. [42]

In the design, all dimensions in the initial filter remained unchanged compare to Figure 3-13 except the length (h) of folded arm in the coupled-line section. In this way, a band-notched UWB filter was implemented and its final layout was shown in the photograph of a fabricated circuit in Figure 6-3. Frequency responses of predicted and measured S_{21} - and S_{11} -magnitudes as well as group delay were illustrated together in Figure 6-4 for comparison. The predicted and measured result had good agreement. In the plotted frequency range of 1.0-12.0 GHz, we can clearly see that the realized in-band graphs that exhibit the emergence of a notch band at 5.59 GHz with 18.8-dB insertion loss and 3-dB notch bandwidth of 4.6%. In the UWB passband except the notch-band, the return loss was higher than 11.0 dB and the group delay was almost flat. Serious variation in group delay was primarily due to the sharp rejection skirt around the notched band.

Figure 6-3. Photograph of an optimized notch band UWB bandpass filter. [42]

Figure 6-4. Predicted & measured S_{21} - and S_{11} - magnitudes and group delay of an optimized notch band UWB bandpass filter. [42]

6.3 Notch-Band Implementation Using Shunt Stubs Loaded Coupled-Line

Most of the reported band notched UWB bandpass filters [33], [35], [68] and [69] suffered from narrow upper-stopband and dull roll-off skirts. In this section, a class of modified UWB bandpass filters are proposed, aiming to simultaneously realize the sharp

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roll-off skirts, wide upper-stopband and controllable notch band at no extra cost. After two types of shunt stubs with symmetrical and asymmetrical configurations were extensively studied, two UWB bandpass filters with or without notch band were in final designed and fabricated on the RT/Duroid 6010 substrate with a dielectric constant of 10.8, loss tangent 0.0023 and a thickness of 0.635 mm. Predicted results with improved UWB filtering performances were experimentally confirmed over the frequency range of 1.0 to 20.0 GHz.

6.3.1 Two-section open-circuited shunt stubs

Figure 6-5. Initial stepped-impedance shunt stub (Stub-A). [41]

Figure 6-5 depicts the layout and circuit model of the initial two-section open-circuited shunt stub, namely, Stub-A. This stub is composed of high-impedance section with Z_1 and low-impedance section with Z_2 that is close to the open-end. For simplification of analysis, these two distinct sections were assumed to have the equalized electrical length of $\lambda_g/4$ at the specified frequency of f_0 that is the central frequency of the desired UWB passband as will be explained later on. As indicated in Figure 6-5, the input impedance Z_{inA} at the intersection point with the two feeding lines can be derived in the closed-form equation (6-

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1) based on the transmission line theorem. With reference to $Z_0=50\ \Omega$ of the two feeding lines, the transmission coefficient, S_{21} , can be obtained and it is shown in equation (6-2).

$$Z_{inA} = jZ_1 \frac{Z_1 \tan(\theta_c) - Z_2 \cot(\theta_c)}{(Z_1 + Z_2)} \quad (6-1)$$

$$S_{21} = \frac{2}{2 - jZ_0 \frac{Z_1 + Z_2}{Z_1(Z_1 \tan(\theta_c) - Z_2 \cot(\theta_c))}} \quad (6-2)$$

where Z_0 , Z_1 and Z_2 are the impedances of two identical feeding line, first and second sections, respectively, while θ_c is the equalized electrical length of the first and second sections in the considered stub, i.e., Stub-A.

Figure 6-6. Folded stepped-impedance shunt stub (Stub-B). [41]

In the following discussion, the frequency (f) is normalized by f_0 , so as to introduce a normalized frequency, $f/f_0=2\theta_c/\pi$. Using the closed-form equation (6-2), S_{21} -magnitude can be easily calculated and Figure 6-7 shows the three curves of its frequency response for three different cases where $Z_2=30, 60$ and $90\ \Omega$ are selected under the fixed $Z_1=90\ \Omega$. As can be observed in Figure 6-7, when Z_2 decreases from $90, 60$ to $30\ \Omega$, the two transmission zeros increasingly move away from the central frequency f_0 . Thus, the 3-dB fractional

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bandwidth (BW) is enlarged from 70%, 90% to 110%. It can be figured out that this fractional bandwidth can be further extended up to 200% as Z_2 is lowered to an infinitely small impedance, i.e., $Z_2=0 \Omega$. In order to sharpen roll-off or rejection skirts in the lower and upper cut-off frequencies of the UWB passband covering 3.1-10.6 GHz, the ratio of these two transmission-zero frequencies need to at least achieve to a value of $10.6/3.1=3.4$. To miniaturize the overall size of these extra stubs, the second section with Z_2 was split into two arms with an impedance of $2Z_2$, thus making up a folded stepped-impedance shunt stub structure, namely, Stub-B, as shown in Figure 6-6. To keep almost no change in filter performances using the above two shunt stub structures, $Z_{inB} \cong Z_{inA}$ needs to be readily set up.

Figure 6-7. Calculated S_{21} -magnitudes of Stub-A and Stub-B using the circuit models. [41]

On the other hand, it is strongly preferred to explore UWB bandpass filters with a single or multiple notch bands [33], [35], [68] and [69] in order not to seriously interfere with the existing wireless systems. To achieve it, a modified shunt stub with asymmetrical geometry

was constructed, namely, Stub-C. Figure 6-8 shows its layout and circuit model. Different from the Stub-B, this modified stub structure is formed by putting these two low-impedance arms of $2Z_2$ in the same side and connecting them together via a short high-impedance section with the length of L as shown in Figure 6-8. The relevant circuit model can be simply constructed by inserting an electrical phase of θ_L to one low-impedance arm in the case of Stub-B.

Figure 6-8. Layout and circuit model of the proposed asymmetrical shunt stub, i.e., Stub-C. [41]

With reference to the work in [35], it can be expected that such a phase difference of θ_L , introduced between the two arms, can create a notch band at a certain frequency (f_z) due to out-of-phase happened in the two arms. To investigate this notch band, equations (6-1) and (6-2) were modified by taking into account an extra phase of θ_L so as to derive equation (6-3) for the input impedance Z_{inC} and equation (6-4) for the transmission coefficient S_{21} . Using equation (6-4), the relationship between the phase shift of θ_L and the normalized transmission-zero frequency of f_z/f_0 can be obtained and Figure 6-9 plots the three relevant graphs under the impedance ratios of $Z_1/Z_2=3.0$, 1.5 and 1.0. As θ_L increases, the notch

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band is gradually shifted down to a lower frequency. Using the graphs in Figure 6-9, the value of θ_L can be straightforwardly determined under the frequency of a desired notch band. In this work, a notch band at 5.6 GHz aims to be produced within the UWB passband with the central frequency at 6.85 GHz. In this case, we can derive $f_z/f_0 = 5.6/6.85 \cong 0.8$. From Figure 6-9, the corresponding θ_L value is found to be almost saturated at $\pi/5$ regardless of varied impedance ratio (Z_1/Z_2)

$$Z_{inC} = jZ_1 \frac{Z_1 + \cot(\theta_c + \theta_L)[Z_1 \tan(\theta_c) - 2Z_2 \cot(\theta_c)]}{Z_1 \cot(\theta_c) + (Z_1 + 2Z_2) \cot(\theta_c + \theta_L)} \quad (6-3)$$

$$S_{21n} = \frac{2jZ_1[Z_1 + \cot(\theta_c + \theta_L)(Z_1 \tan(\theta_c) - 2Z_2 \cot(\theta_c))]}{2jZ_1[Z_1 + \cot(\theta_c + \theta_L)(Z_1 \tan(\theta_c) - 2Z_2 \cot(\theta_c))] + Z_o[Z_1 \cot(\theta_c) + (Z_1 + 2Z_2) \cot(\theta_c + \theta_L)]} \quad (6-4)$$

Figure 6-9. Electrical length (θ_L) versus normalized transmission-zero frequency (f_z/f_0). [41]

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Figure 6-10. Calculated S_{21} -magnitudes with notch band. [41]

To view the whole frequency response of the shunt stub in Figure 6-8 with different Z_1/Z_2 , Figure 6-10 plots the three graphs of S_{21} -magnitude versus $Z_2 = 90, 60$ and 30Ω under the fixed $\theta_L = \pi/5$. On one hand, a notch band is really created at the fixed frequency, i.e., $f_z/f_0 = 0.8$. On the other hand, the spacing between the lower and upper transmission zeros becomes wide with respect to Z_2 . Its variation rarely receives significant influence as the notch-band is introduced, therefore the two transmission zeros can still be freely adjusted according to the specified bandwidth of the main passband. It is worthwhile to note that, for $\pi > \theta_L > 0$, the notch band is located within the range of $0 < f_z/f_0 < 1$. However, after our extensive investigation on the asymmetrical structure in Figure 6-8, we can ignore the two extreme cases described as below. If a notch band was in the range of $f_z/f_0 < 0.4$, the lower transmission zero would be “pushed out” from the concern frequency range $0 < f/f_0 < 1$ and the notch band frequency will be located in the out-of-band. When $\theta_L = 0$, the lengths of two split stubs become the same. The circuit model of Stub-C becomes exactly the same as

that in Stub-B, so the notch band will be fully disappeared. Using the similar approach, one can design the notch band above the center frequency $f/f_0=1.0$ by forcing the electrical length of $\theta_L+\theta_c<\lambda_g/4$.

6.3.2 Parallel coupled-lines with shunt stubs

Figure 6-11. Layout and circuit model of UWB bandpass filter based on coupled line loaded with two symmetrical Stub-B structures, namely, Type-A. [41]

Now, let’s look at Equations (6-3) and (6-4) again, in which there are a few assumptions described as below for simplification of analysis. The electrical lengths of the first and second sections in a symmetrical geometry have the same value, i.e., θ_C ; the impedance of the short connecting line of θ_L is assumed to be $2Z_2$; there exists no coupling between the two parallel portions in the Stub-B. As it has been extensively studied, equivalent circuit models provide us with a simple physic-based approach in the initial design stage. Due to several unexpected discontinuity effects such as step, bend, coupled interaction and frequency dispersion, the results based on circuit models are approximate,

but they can be used as the initial parameters when the accurate optimization design is executed based on the time-consuming full-wave simulation tools/approaches.

Figure 6-12. Simulated S_{21} - and S_{11} -magnitudes of Type-A UWB bandpass filter based on coupled line loaded with two symmetrical Stub-B structures. [41]

In the following, the above-investigated Stub-B and Stub-C structures are used to make up two initial types of UWB bandpass filters with or without notch band, namely, Type-A and Type-B. Figure 6-11 shows the layout of the first initial UWB bandpass filter and its equivalent circuit model. The parallel coupled microstrip line with one quarter wavelength ($\lambda_g/4$ at 6.85 GHz) can create a wide passband ranging from the DC to the frequency of one half-wavelength. The two Stub-B structures are shunt-connected with the two individual strip conductors before their coupling portion. In the circuit model, the parallel coupled-line section is modeled as an equivalent J-inverter network as shown in the center part of Figure 6-11. This circuit model gives us a qualitative description on its operating mechanism and allows us to select the initial dimensions of its overall structure for full-wave simulation

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and design. Figure 6-12 shows the simulated frequency responses of S_{21} - and S_{11} -magnitudes based on the full-wave Agilent Momentum software. As discussed in Section 6.3.1, the Stub-B itself can produce two transmission zeros at lower and upper stopbands. It works together with the coupled-line so as to realize a UWB passband performance with sharp roll-off rejections. This UWB bandpass filter has the return loss in the passband higher than 20 dB. Nevertheless, the unwanted passbands or harmonics unexpectedly appear at frequencies above 16 GHz, thereby causing a poor out-of-band problem.

Figure 6-13. Layout and circuit model of UWB bandpass filter based on coupled line loaded with two asymmetrical Stub-B structures, namely, Type-B. [6]

Figure 6-13 depicts the circuit layout and its equivalent circuit model of a notch-band UWB bandpass filter. Instead of Stub-B, the two Stub-C structures are utilized herein. Simulated frequency responses of S_{21} - and S_{11} -magnitudes are plotted in Figure 6-14. This filter still has the good passband with sharp rejection skirt except that a notch band is

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produced at the desired frequency of 5.6 GHz. As a trade-off of this notch band, the upper-stopband gets somehow worse due to an extra section of L in the Stub-C. To address this issue of poor upper-stopband, an EBG-embedded resonator structure [38] is further utilized in the next section with two stub-loaded coupled-lines.

Figure 6-14. Simulated S_{21} - and S_{11} -magnitudes of Type-B UWB bandpass filter based on coupled line loaded with two asymmetrical Stub-B structures. [41]

6.3.3 Experimental result and discussions

Figure 6-15. layout and dimensions of proposed UWB bandpass filter with notch band. [41]

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Figure 6-16. Complete circuit model of proposed UWB bandpass filter with notch band. [41]

In this section, the above two types of initial UWB filters based on a single stub-loaded coupled-line were cascaded via EBG transmission line, aiming to build up the two UWB microstrip-line bandpass filters with and without notch band. From our alternative view as described in [38], the center EBG line works together with two high-impedance sections at two sides to make up an upper-stopband included multiple-mode resonator. In this way, such an ultra-wide passband can be simply interpreted based on simultaneous occurrence of the first three resonant modes. Figure 6-15 illustrates the layout of a notch-band UWB bandpass filter and its relevant complete circuit model is shown in Figure 6-16. By replacing the two Stub-C with the two Stub-B, its corresponding UWB bandpass filter without notch band can be straightforwardly formulated. In the design, the identical shunt stubs, i.e., Stub-B or Stub-C, with coupled-lines were cascaded together via EBG line in the middle. Then, the EBG line was modeled with the target that the three resonant modes of the whole EBG-embedded resonator equally occur within the UWB passband while the bandstop was produced above 16 GHz. In final, these UWB filters are optimally designed towards high in-band and low out-of-band return losses using the Momentum software.

Figure 6-17. Top-view photograph of the notch-band excluded UWB bandpass filter, namely, Type-C. [41]

Figure 6-17 and 6-19 depict the photographs of the two fabricated UWB bandpass filters with and without notch band. Their overall lengths are 14.5 mm and 12.5 mm, respectively. Figure 6-18 is the predicted frequency responses of S_{21} - and S_{11} -magnitudes as well as group delays of the Stub-B loaded UWB filter without notch band. As can be observed in Figure 6-18, the dominant passband satisfactorily achieves the specified UWB of 3.1-10.6 GHz with the return loss higher than 11.0 dB. Plus, the insertion loss in the upper-stopband is higher than 20.0 dB in the upper-stopband of 11.0-20.0 GHz. These results prove that the proposed Stub-B loaded UWB bandpass filter can not only extend the upper-stopband with increased attenuation but also largely raise the preferred roll-off skirts in both lower and upper cut-off frequency edges. The group delay is shown in the Figure 6-18 in the range of 2.0-12.0-GHz. The measured group delay is well matched with its predicted counterpart until the frequency of 10.6 GHz. At or beyond 10.6 GHz, the S_{21} -magnitude is kept lower than -20 dB as shown Figure 6-18. Therefore, it is very difficult in theory to accurately extract the S_{21} -phase at high frequencies due to extremely small S_{21} -magnitude and too long electrical feed line. Within the UWB passband, the group delay at the mid-frequency is about 0.2 ns and its shape is very flat as the frequency varies in the middle range. Thus, this UWB filter (Type-C) has a good linearity in the UWB passband.

Figure 6-18. S_{21} - and S_{11} -magnitudes of predicted and measured frequency responses of the Type-C notch-band excluded UWB bandpass filter. [41]

Figure 6-20 is the predicted and measured frequency responses of S_{21} - and S_{11} -magnitudes as well as group delay of the Stub-C loaded UWB bandpass filter with notch band (Type-D) as shown in Figure 6-19. They are found to be in very reasonable agreement with each other over the plotted range. The insertion loss of the upper-stopband is higher than 16 dB in the 10.8-20.0-GHz frequency range. The roll-off skirts are as sharp as those in Figure 6-18 and the notch band finds to be well located at the specified 5.6 GHz with 16 dB attenuation in measurement. Comparing Figure 6-20 with Figure 6-18, the group delay is 0.2 ns at the center frequency and it is almost the same as each other except at 5.6 GHz (notch-band). The serious variation of this group delay near the notch band is primarily due to the resonance of the added stub-C. Since the signal in this notch band is highly attenuated, such a variation in group delay hardly distorts the signal waveform.

Figure 6-19. Top-view photograph of the notch-band included UWB bandpass filter, namely, Type-D. [41]

Figure 6-20. S_{21} - and S_{11} -magnitudes of the notch-band included Type-D UWB bandpass filter, namely, Type-D. [41]

6.4 Conclusion

In this Chapter, two approaches were introduced to design notch-band UWB bandpass filter. Firstly, a notch-band filter based on previous introduced UWB filter was implemented by modifying the arms of interdigital coupled-line to asymmetrical. The

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measurement results matched with the simulation result. This notch-band approach is applicable to all the UWB filters that had been discussed in previous chapters with interdigital coupled-line.

Later on, three types of open-circuited shunt stubs with varied configurations, i.e. Stub-A, Stub-B and Stub-C, were at first studied to exhibit their distinctive transmission-zero properties. Based on the simple equivalent circuit models, the two sets of closed-form equations were obtained to estimate the locations of these transmission zeros as verified by full-wave-simulation. The Stub-B and Stub-C were then shunt-connected with a parallel coupled-line with high impedance to derive the initial UWB passband performance due to reallocations of low and higher transmission zeros. Next, the EBG transmission line was sandwiched between the identical stub-loaded parallel coupled-lines, so as to improve the upper-stopband to a large extent. The two designed UWB bandpass filters with and without notch band were finally designed, fabricated and measured. All the filters achieve sharp roll-off skirts at the cut-off frequency edges and wide upper-stopband beyond 20 GHz. It was confirmed in the experiment that the filters with symmetrical stubs (Stub-B) and asymmetrical stubs (Stub-C) had really achieved good UWB in-band performances with and without a notch band at 5.6 GHz within the realized UWB passband.

Chapter 7

UWB Power Dividers

7.1 Introduction

Power dividers are one of the key passive components in many microwave circuits and subsystems. The most famous and useful power divider is the Wilkinson power divider that has excellent isolation between two output ports [70]. However it has less than 20% fractional bandwidth. Due to the rapid growth of unlicensed use of ultra-wideband (UWB) for short-distance wireless communications, there has been tremendous interest in exploration of various UWB components that operate in the UWB range, i.e., 3.1-10.6 GHz. For this purpose, a few wideband power dividers [71]-[76] have been developed so far. Normally, wideband divider can be constituted by cascading multi-section matching networks at two output ports of a single Wilkinson power divider. Nevertheless, this approach inquires more resistors for isolation. According to the analysis in [71], at least three sections and three resistors are needed to achieve the desired ultra-wideband frequency response. In [72], a multi-section impedance transformer was placed at the input port of a multi-way junction to obtain good wideband return loss, but this divider suffered

from poor isolation due to the absence of any resistor between two outputs. In [73], three stage wideband Wilkinson power divider was explored for differential application. In [5], three stage wideband Wilkinson power divider was developed with an arbitrary power division ratio between two output ports. Recently, special attention has been shifted to exploration of ultra-wideband power dividers with the operating band in the overall 3.1-10.6-GHz UWB region. In [75], an out-of-phase UWB power divider is implemented with microstrip to slotline transition. In [76], a wideband divider is constituted with broadside coupling structure.

In this Chapter, three UWB power dividers on single-layer microstrip-line are proposed, analyzed and designed. Only a single resistor is needed for isolation purpose among these UWB power dividers. Thus, good power splitting, impedance matching and isolation can be achieved over the UWB range. In the following, the proposed UWB dividers will be designed on the RT/Duroid 6010 substrate with a dielectric constant of 10.8 and thickness of 0.635 mm.

7.2 Power Divider with Pairs of Shunt Short-Circuit Stubs

The schematic of the first proposed power divider, namely, Type-A, is shown in Figure 7-1:

Figure 7-1. The schematic of proposed Type-A UWB microstrip power divider. [79]

It is different from the traditional multi-stage Wilkinson power divider [73]. Two pairs of quarter-wavelength short-circuited stubs were symmetrically placed at the two outputs and they were joined through a pair of quarter-wavelength impedance transformer. Effectively, each arm of the divider included a circuit that was similar to the two-port UWB bandpass filter with short-circuited stubs [32]. In principle, it is thus possible to achieve an ultra-wideband power splitting performance by properly designing the two individual impedance transformers with the characteristic impedance of Z_1 and Z_2 as well as the two identical short-circuited stubs with the characteristic impedance of Z_3 . Moreover, different from the two existing UWB power dividers [75]-[76] with poor isolation, a single resistor R was placed between two output parts and its value was properly selected to achieve good isolation between port 2 and port 3 over the required UWB range.

7.2.1 Odd-mode analysis

Figure 7-2. The odd-mode model of the proposed Type-A UWB microstrip power divider.[79]

Figure 7-2 shows the two-port transmission-line model for a symmetrical bisection of the UWB power divider in Figure 7-1 under odd-mode excitation. In this case, port 1 can simply form a short circuit since the strip conductor of this port is grounded by the virtual electric wall. The model becomes a one-port network as shown in Figure 7-2, and the input admittance at port 2, can be derived as

$$Y_{in2}^o = Y_2 \frac{Y_1^o + jY_2 \tan(\theta)}{Y_2 + jY_1^o \tan(\theta)} - jY_3 \cot(\theta) \quad (7-1)$$

where $Y_1^o = 2/R - jY_3 \cot(\theta) - jY_1 \cot(\theta)$.

As illustrated in Figure 7-2, $Z_{p2} = 1/Y_{p2} = 50 \Omega$ represents the port impedance at port 2 of the odd-mode bisection. At the center frequency of UWB, i.e., 6.85 GHz, $\theta = \pi/2$ and $Y_{in2}^o = Y_{p2}$ implies no reflection at port 2. Thereafter, equation (7-1) can be simplified as

$$R = \frac{Z_2^2}{0.5Z_{p2}} \quad (7-2)$$

Equation (7-2) gives a simple guideline in selection of R . For instance, if $Z_2 = 50 \Omega$, $R = 100$

Ω_4 is obtained.

7.2.2 Even-mode analysis

Figure 7-3. The even-mode model of the proposed Type-A UWB microstrip power divider. [79]

Under the even-mode excitation, a symmetrical bisection of the three-port divider can be represented by a two-port network shown in Figure 7-3. Herein, the center vertical plane becomes perfect magnetic wall. The port impedance at port 1 is doubled, i.e., $Z_{p1}=100\ \Omega$, since its relevant strip conductor is cut into halves. For the two-port network in Figure 7-3, the input admittances at port 1 and port 2 can be derived on a basis of the transmission-line theory and they are expressed in equations (7-3) and (7-4).

$$Y_{in1}^e = Y_1 \frac{Y_{t1}^e + jY_1 \tan(\theta)}{Y_1 + jY_{t1}^e \tan(\theta)} \quad (7-3)$$

$$Y_{in2}^e = Y_2 \frac{Y_{t2}^e + jY_2 \tan(\theta)}{Y_2 + jY_{t2}^e \tan(\theta)} - jY_3 \cot(\theta) \quad (7-4)$$

where

$$Y_{i1}^e = Y_2[Y_i^e + jY_2 \tan(\theta)]/[Y_2 + jY_i^e \tan(\theta)] - jY_3 \cot(\theta)$$

$$Y_i^e = Y_{p2} - jY_3 \cot(\theta)$$

$$Y_{i2}^e = Y_1[Y_{p1} + jY_1 \tan(\theta)]/[Y_1 + jY_{p1} \tan(\theta)] - jY_3 \cot(\theta)$$

To achieve good impedance matching at port 1 and port 2, $Y_{in1}^e = Y_{p1}$ and $Y_{in2}^e = Y_{p2}$ need to be restricted to an acceptable extent. Under this constraint, one can simplify equations (7-3) and (7-4) as a unified expression at the frequency of $\theta = \pi/2$.

$$Z_1^2 = \frac{Z_2^2 Z_{p1}}{Z_{p2}} \quad (7-5)$$

Due to $Z_{p1} = 100 \Omega$ and $Z_{p2} = Z_2 = 50 \Omega$ in this even-mode case, one can determine $Z_1 = 70.7 \Omega$ from (7-5). It should be emphasized that Equations (7-2) and (7-5) are only exactly valid at $\theta = \pi/2$ and impedance-matching may be gradually degraded as θ deviates away from $\pi/2$. It is the same case as the traditional Wilkinson power divider [70]. Nevertheless, when short-circuited stubs are placed as shown in Figure 7-1, two additional transmission poles can be generated at the two sides of $\theta = \pi/2$. Good impedance matching can be achieved in the overall UWB region if these three poles are properly allocated as discussed in [32].

7.2.3 S-parameter analysis

As the input admittances at port 2 and port 3 are derived in (7-1), (7-3) and (7-4) for even- and odd-mode cases, the resultant S-parameters, S_{11} and S_{21} , can be expressed as,

$$S_{11} = \frac{Y_{p1} - Y_{in1}^e}{Y_{p1} + Y_{in1}^e} \quad (7-6)$$

$$S_{22} = \frac{1}{2} \left(\frac{Y_{p2} - Y_{in2}^e}{Y_{p2} + Y_{in2}^e} + \frac{Y_{p2} - Y_{in2}^o}{Y_{p2} + Y_{in2}^o} \right) \quad (7-7)$$

With these two closed-form equations, frequency response of S-parameters can be calculated for the desired frequency range. Figure 7-4 and Figure 7-5 shows the calculated S-parameter magnitudes, i.e., $|S_{11}|$, $|S_{22}|$, $|S_{21}|$ and $|S_{32}|$ of the three-port divider in Figure 7-1 using the impedance of $Z_3=90\ \Omega$ and varied values of $Z_2=40, 50$ and $60\ \Omega$. Comparing the cases of $Z_2=40$ and $60\ \Omega$ in Figure 7-4, the selection of $Z_2=50\ \Omega$ provides a better matching at port 1 over a very wide bandwidth. As shown in Figure 7-5, the selection of $Z_2=50\ \Omega$ achieves a good isolation and matching at port 2. Thus, we can conclude that $Z_2=50\ \Omega$ is the best value of them for this power divider.

Figure 7-4. Simulated S-parameters with respect to electrical length (θ) for the Type-A UWB power divider. [79]

Figure 7-5. Simulated S-parameters with respect to electrical length (θ) for the Type-A UWB power divider. [79]

In this case, the fractional bandwidths under the definitions of 3-dB $|S_{11}|$ and $|S_{22}|$ are found to be about 129% and 121%, respectively. Meanwhile, we can find that $|S_{21}|$ achieves very flat 3-dB power splitting performance in the frequency band of 121% fractional bandwidth and sharp roll-off skirt near the lower and upper cut-off frequencies. This excellent transmission performance with 3-dB return loss in the UWB is mainly brought out by the two in-band transmission poles of two pairs of short-circuited stubs. Different from [75]-[76], $|S_{32}|$ in Figure 7-5 is lower than -13 dB over the plotted frequency range as the resistor of $R=100\ \Omega$ is introduced. Thus, this designed power divider not only shows excellent return and insertion losses but also attains reasonably good isolation in the overall UWB band with the fractional bandwidth of 121%.

Figure 7-6. Electrical length (θ) of lower and upper cut-off frequencies versus characteristic impedance (Z_3) of short-circuited stubs for the Type-A UWB power dividers. [79]

To comparatively investigate their ultra-wideband power splitting performances, the lower/upper cut-off frequencies can be calculated via Equations (7-6) and (7-7) under the 3-dB bandwidth definition for $|S_{11}|$ and $|S_{22}|$. Figure 7-5 illustrates the calculated electrical lengths or phases (θ) at the two cut-off frequencies versus characteristic impedance (Z_3) of short-circuited stubs for the Type-A divider. As discussion in Figure 7-4, the fractional bandwidths of these two power dividers can be expressed as the difference of the upper/lower phases (θ) over its central value since θ is linearly proportional to the frequency (f).

With reference to Figure 7-6, these two cut-off phases (θ) are found to go up and down simultaneously as Z_3 increases from 40 to 200 Ω , progressively enlarging the spacing between them or widening the operating frequency bandwidth. Therefore the bandwidth of the Type-A divider can be easily varied with virtue of adjustment of Z_3 . In Figure 7-6, the dotted straight line (a-a') represents the case of $Z_3=90 \Omega$ (Figure 7-4 and Figure 7-5) and its

interception points with two sets of θ -curves allows us to calculate the fractional bandwidths, i.e., 129% and 121% for $|S_{11}|$ and $|S_{22}|$.

7.3 Power Divider with Coupled-Line Section

Figure 7-7. The schematic of proposed Type-B UWB microstrip power divider with coupled-line.

In this section, an alternative UWB power divider, namely, Type-B, was constituted with the target of improved out-of-band performance. Figure 7-7 shows its schematic. In comparison to the initial UWB divider in Figure 7-1, a pair of high-impedance coupled-lines replaces a pair of impedance transformers in this Type-B divider. As discussed in [16], the introduction of these coupled-lines renders two additional transmission poles in the UWB. In principle, the roll-off skirts near the lower and upper cut-off frequencies can be effectively sharpened since the total number of transmission poles is increased inside the

UWB. For analysis, the two modified equivalent models were formulated under the odd- and even- mode excitations. Figure 7-8 and 7-9 depict these models, where the equivalent model of coupled-line section was derived based on the coupled-line theory [77]. Proceeding with the similar analysis in Section II, equations (7-1)-(7-5) can be modified as (7-8)-(7-12) imposing the constraints of 3-dB $|S_{21}|$ as well as minimum $|S_{11}|$, $|S_{22}|$ and $|S_{32}|$.

Figure 7-8. The odd-mode model of proposed Type-B UWB microstrip power divider with coupled-line.

In the odd-mode case, the input admittance at port 2 in Figure 7-7 can be derived as

$$Y_{in2}^o = \frac{4j + 2Y_t^o Z_+ \cot(\theta)}{2Z_+ \cot(\theta) + jY_t^o Z_-^2 \csc^2(\theta) - jY_t^o Z_+^2 \cot^2(\theta)} - jY_3 \cot(\theta) \quad (7-8)$$

where $Y_t^o = R/2 - jY_3 \cot(\theta) - jY_1 \cot(\theta)$, $Z_- = Z_{oe} - Z_{oo}$ and $Z_+ = Z_{oe} + Z_{oo}$. Z_{oe} and Z_{oo} are the even- and odd-mode characteristic impedances of the coupled-line section, respectively.

Figure 7-9. The even-mode model of proposed Type-B UWB microstrip power divider with coupled-line.

Based on the equivalent even-mode network in Figure 7-8, the input admittance at port 1 can be obtained

$$Y_{in1}^e = Y_1 \frac{Y_{t1}^e + jY_1 \tan(\theta)}{Y_1 + jY_{t1}^e \tan(\theta)} \quad (7-9)$$

where

$$Y_{t1}^e = \frac{4j + 2Y_t^e Z_+ \cot(\theta)}{2Z_+ \cot(\theta) + jY_t^e Z_-^2 \csc^2(\theta) - jY_t^e Z_+^2 \cot^2(\theta)} - jY_3 \cot(\theta)$$

$$Y_t^e = Y_{p2} - jY_3 \cot(\theta)$$

and the input admittance at port 2 becomes

$$Y_{in2}^e = \frac{4j + 2Y_{t2}^e Z_+ \cot(\theta)}{2Z_+ \cot(\theta) + jY_{t2}^e Z_-^2 \csc^2(\theta) - jY_{t2}^e Z_+^2 \cot^2(\theta)} - jY_3 \cot(\theta) \quad (7-10)$$

where

$$Y_{t2}^e = Y_1 \frac{Y_{p1} + jY_1 \tan(\theta)}{Y_1 + jY_{p1} \tan(\theta)} - jY_3 \cot(\theta)$$

Similar to the analysis approach in previous section, the good impedance matching constraint at all the three ports at the frequency of $\theta = \pi/2$ leads to derivative of the following two closed-form formulae.

$$R = \frac{Z_-^2}{2Z_{p2}} \quad (7-11)$$

$$Z_1^2 = \frac{Z_{p1}Z_-^2}{4Z_{p2}} \quad (7-12)$$

It should be noted that both two-way power dividers have $Z_{p2}=Z_{p3}=50 \Omega$. Based on (7-11), if $R=100 \Omega$ is selected, we can obtain $Z_-=100 \Omega$, where $Z_{oe}=135 \Omega$ and $Z_{oo}=35 \Omega$. Similar to the Type-A divider, the port impedances at port 1 and port 2 still become $Z_{p1}=Z_-=100 \Omega$ and $Z_{p2}=50 \Omega$. Thus, from (7-12), we can calculate $Z_1=70.7 \Omega$, under the constraint of good return/insertion losses, $Z_3=60 \Omega$ for the short-circuited stubs can be determined.

Figure 7-10. Electrical length (θ) of lower and upper cut-off frequencies versus characteristic impedance (Z_3) of short-circuited stubs for the Type-B UWB power dividers.

Figure 7-11. Electrical length (θ) of lower and upper cut-off frequencies versus characteristic impedance (Z_3) of short-circuited stubs for the Type-B UWB power dividers.

Figure 7-10 and Figure 7-11 plot the calculated S-parameter magnitudes over a wide range of electrical length (θ) that is quasi-linearly proportional to frequency. From equations (7-11) and (7-12), we can derive a matching condition, i.e., $Z=100\ \Omega$, where $Z=Z_{oe}-Z_{oo}$. In order to obtain a set of optimized value of Z_{oe} and Z_{oo} , three sets of Z_{oe} and Z_{oo} are plotted in Figure 7-10 and Figure 7-11. As can be seen in Figure 7-10, the set with $Z_{oe}=155\ \Omega$ and $Z_{oo}=55\ \Omega$ has narrow bandwidth matching at port 1 and some undesired $|S_{21}|$ spurious appear near the cut-off region. The set with $Z_{oe}=115\ \Omega$ and $Z_{oo}=15\ \Omega$ has very good matching at port 1 and sharp rejection skirt for the $|S_{21}|$, but the bandwidth is not wide enough to cover the desired passband. Comparatively, the set with $Z_{oe}=135\ \Omega$ and $Z_{oo}=35\ \Omega$ can be considered as the best values as it has good wideband matching at port 1 with return loss larger than 10 dB and sharp rejection at the cut-off region. Now let's look at Figure 7-11, the isolation $|S_{23}|$ change slightly when the three sets of the Z_{oe} and Z_{oo} are considered.

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However, the set with $Z_{oe}=135\ \Omega$ and $Z_{oo}=35\ \Omega$ has wider bandwidth for the port 2 matching when compared with other two sets of Z_{oe} and Z_{oo} . It is concluded that the set with $Z_{oe}=135\ \Omega$ and $Z_{oo}=35\ \Omega$ is the best values for this power divider. Since the coupled-line generated two additional transmission zeros at $\theta=0$ and π [16]. The roll-off skirt of Type-B divider is obviously sharper than that of the Type-A counterpart if two sets of graphs in Figure 7-4 and Figure 7-10 are compared (solid lines). From Figure 7-10, we can find that the fractional bandwidth is 135% which is slightly wider than the 121% bandwidth observed in case of Type-A divider. Moreover, we can see that all of $|S_{11}|$, $|S_{22}|$ and $|S_{32}|$ are lower than -10 dB and $|S_{21}|$ is close to -3 dB over the desired UWB passband.

Figure 7-12. Electrical length (θ) of lower and upper cut-off frequencies versus characteristic impedance (Z_3) of short-circuited stubs for the Type-B UWB power dividers.

Now, let's consider the bandwidth for the Type-B divider. In Figure 7-12, the bandwidth for $|S_{11}|$ is still wider than that for $|S_{22}|$ in this coupled-line case. Due to the presence of two additional transmission poles in this case, its upper and lower cut-off

phases (θ) would become larger and smaller than those of its counterpart Type-A if the impedance Z_3 was unchanged. The fractional bandwidth of this Type-B divider can be further enlarged if necessary. With the target of the same bandwidth as that in the Type-A, we can determine $Z_3=60\ \Omega$ that is represented by the b-b' dotted-line in Figure 7-12. The S-parameters results from this b-b' line can refer to Figure 7-10 and Figure 7-11.

7.4 Power Divider with Coupled-Line Section and Shunt Stepped-Impedance Stubs

Figure 7-13. Schematic of the proposed Type-C UWB power divider on microstrip line. [78]

The schematic of the proposed shunt stepped-impedance UWB power divider named Type-C is shown in Figure 7-13. As a signal is incident at port 1, its power can be equally divided to port 2 and port 3 over a wide frequency band. For the purpose of good out-of-band rejection and good in-band matching, a pair of parallel coupled-lines and stepped-impedance open-circuited stubs was introduced at two output ports of a traditional Wilkinson power divider. Following the basic idea in the UWB bandpass filters in [16], [32], multiple transmission poles can be produced over the UWB. Further, to prevent signal transmission between port 2 and port 3, a single resistor of R was installed. Based on the even- and odd-mode analysis [77], its UWB characteristics can be analyzed in terms of S-parameters.

Figure 7-14. The odd-mode circuit model of the proposed UWB power divider on microstrip line. [78]

Figure 7-14 shows the two-port circuit model for a bisection of the Type-C UWB power divider in Figure 7-13 under odd-mode excitation. In this case, port 1 can be simply considered as a short circuit since the strip conductor of this port is grounded via virtual

electric wall. Thus, this model becomes a one port network and the input impedance at port 2 is derived as:

$$Z_{in2}^o = \frac{2Z_+ \cot(\theta) + jY_l^o Z_-^2 \csc^2(\theta) - jY_l^o Z_+^2 \cot^2(\theta)}{4j + 2Y_l^o Z_+ \cot(\theta)} \quad (7-13)$$

where, $Y_l^o = 2/R - jY_1 \cot(\theta) + 1/Z_{inA}$, $Z_- = Z_{oe} - Z_{oo}$, $Z_+ = Z_{oe} + Z_{oo}$ and $Z_{inA} = jZ_2(Z_2 \tan(\theta) - Z_3 \cot(\theta))/(Z_2 + Z_3)$. Z_{oe} and Z_{oo} are the even- and odd-mode impedances of the coupled-line respectively.

In Figure 7-14 and 7-15, Z_{p1} and Z_{p2} represent characteristic impedance at port 1 and 2 respectively. Under the constraint at port 2 of $Z_{in2}^o = Z_{p2} = 50 \Omega$ at $\theta = \pi/2$, equation (7-13) can be simplified as

$$R = \frac{Z_-^2}{2Z_{p2}} \quad (7-14)$$

and this provides a simple guideline in selection of R . For instance, if $R = 100 \Omega$ is selected, $Z_- = 100 \Omega$ is obtained.

Figure 7-15. The even-mode circuit model of the proposed UWB power divider on microstrip line. [78]

Under the even-mode excitation, a bisection of the divider in Figure 7-13 is expressed as a two-port network shown in Figure 7-15. The symmetrical plane becomes a perfect magnetic wall and the characteristic impedance at port 1 is double Z_{p1} , or 100 Ω . Thus, the input impedances at port 1 and port 2 can be derived as:

$$Z_{in1}^e = Z_1 \frac{1 + jY_{i1}^e Z_1 \tan(\theta)}{Y_{i1}^e Z_1 + j \tan(\theta)} \quad (7-15)$$

$$Z_{in2}^e = \frac{2Z_+ \cot(\theta) + jY_{i2}^e Z_-^2 \csc^2(\theta) - jY_{i2}^e Z_+^2 \cot^2(\theta)}{4j + 2Y_{i2}^e Z_+ \cot(\theta)} \quad (7-16)$$

where

$$Y_{i1}^e = \frac{4jZ_{p2} + 2Z_+ \cot(\theta)}{2Z_{p2}Z_+ \cot(\theta) + jZ_-^2 \csc^2(\theta) - jZ_+^2 \cot^2(\theta)} + \frac{1}{Z_{in4}}$$

$$Y_{i2}^e = [Z_1 + jZ_{p1} \tan(\theta)] / [Z_1(Z_{p1} + jZ_1 \tan(\theta))] + 1 / Z_{in4}$$

Under the constraint of good impedance matching at port 1 and port 2, i.e., $Z_{in1}^e = Z_{p1}$ and $Z_{in2}^e = Z_{p2}$, one can simplify equation (7-15) and (7-16) as a unified expression at the frequency of $\theta = \pi/2$.

$$Z_1^2 = \frac{Z_{p1}Z_-^2}{4Z_{p2}} \quad (7-17)$$

Due to $Z_{p1} = Z_- = 100 \Omega$ and $Z_{p2} = 50 \Omega$, one can determine $Z_1 = 70.7 \Omega$ from (7-17). Noted that (7-14) and (7-17) are only accurate at $\theta = \pi/2$ and they are gradually degraded as θ

deviates away from $\pi/2$. However, as stepped-impedance stubs and coupled-lines are installed as illustrated in Figure 7-13, four additional transmission poles can be generated, two on each sides of $\theta=\pi/2$. Thus, good impedance matching still can be achieved in the overall UWB range if these poles are properly allocated.

Figure 7-16. Electrical length (θ) of open-circuited stubs at the lower/upper 3-dB cut-off frequencies against impedance ratio (Z_2/Z_3) for Type-C UWB power divider. [78]

Figure 7-17. S-parameters of the proposed Type-C UWB power divider with stepped-impedance stubs. [78]

The three individual input impedances are derived in (7-13), (7-15) and (7-16). Thus, S_{11} and S_{22} can be simply calculated. In order to demonstrate the performance of this Type-C UWB power divider, the lower/upper cut-off frequencies of the passband can be derived under the definitions of 3-dB S_{11} - and S_{22} -magnitudes. Figure 7-16 plots the calculated electrical length (θ) at the two cut-off frequencies versus impedance ratio (Z_2/Z_3) of the stepped- impedance stubs. These two cut-off frequencies are found to go up and down respectively as the ratio (Z_2/Z_3) increases from 0.6 to 9.0. Figure 7-14 is plotted to determine the operation bandwidth of the divider and the vertical plane $Z_2/Z_3=2.25$ (a-a') in Figure 7-16 achieves the 110% fractional bandwidth that corresponds to the 3.1-to-10.6-GHz UWB. Then, $Z_3=40\ \Omega$ if $Z_2=90\ \Omega$ is chosen. The four independent S-parameters, i.e., $|S_{11}|$, $|S_{21}|$, $|S_{22}|$ and $|S_{32}|$, can be calculated and they are plotted over a wide range of the phase (θ) in Figure 7-17. In addition to good in-band splitting, impedance matching and isolation performances, Figure 7-17 demonstrates that this Type-C UWB power divider can achieve much sharper roll-off skirts near the lower/upper cut-off frequencies than the Type-A and Type-B. It is mainly benefited from the transmission zeros of the stepped-impedance stubs and coupled lines. $|S_{11}|$ and $|S_{22}|$ are lower than -13 dB and -20 dB respectively while $|S_{32}|$ between two output ports is lower than -10 dB over the UWB passband.

7.5 Discussion and Experimental Verification

As the three UWB dividers were designed in Sections 7.2, 7.3 and 7.4, three prototype UWB dividers were fabricated and measured. Figure 7-18 is the photographs of the fabricated Type-B UWB power divider with the three SMA connectors at three ports

respectively:

Figure 7-18. Photograph of the Type-B UWB power divider. [80]

Figure 7-19 and 7-20 shows the predicted S_{21} - and S_{11} -magnitudes as well as the group delay in comparison to the measured results for the Type-A and Type-B respectively. They are in reasonably good agreement with each other over the frequency range from 1 to 14 GHz, exhibiting good performance in terms of 3-dB power splitting. The return loss is higher than 10 dB over the desired passband and its roll-off skirt near the lower and cut-off frequencies, i.e., 3.1 and 10.6 GHz, has been effectively sharpened as expected in theory. The spurious passband is observed in the frequencies higher than 16.5 GHz and it is primarily due to the periodic harmonics of coupled-lines and short-circuited stubs as also existed in [8]-[9]. The group delay of $|S_{21}|$ is very flat over the passband for both Type-A and Type-B. This implies good linearity of this power divider within the operation frequency range. Figure 7-20 and 7-21 illustrates the comparison between the predicted and measured S_{22} - and S_{32} -magnitudes. Again, they are in good agreement with each other. The return loss at port 2 and port 3 as well as the isolation between port 2 and port 3 are all

higher than 10 dB over the realized UWB band.

Figure 7-19. Predicted and measured results S_{21} - and S_{11} -magnitudes as well as group delay of the designed Type-A UWB power divider. [79]

Figure 7-20. Predicted and measured results S_{22} - and S_{32} -magnitudes of the designed Type-A UWB power divider. [79]

Figure 7-21. Predicted and measured results S_{21} - and S_{11} -magnitudes as well as group delay of the designed Type-B UWB power divider. [80]

Figure 7-22. Predicted and measured results S_{22} - and S_{32} -magnitudes of the designed Type-B UWB power divider. [80]

Figure 7-23 shows photograph of Type-C UWB power divider. Figure 7-24 plots the predicted and measured $|S_{11}|$ and $|S_{21}|$ in a frequency range of 1.0 to 14.0 GHz as well as the group delay. They matched with each other. The measured $|S_{11}|$ is lower than -10.0 dB and

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$|S_{21}|$ is close to -3.0 dB over the 3.5 to 10.1 GHz with 0.4 dB insertion loss at center frequency in measurement. In particular, the roll-off skirts near the two cut-off frequencies are significantly sharpened as demonstrated in simulation. And also, the group delay shows good linearity within the passband. There are discrepancies between the measured and simulated result in $|S_{21}|$ especially at higher frequency, this is due to the small gap (0.06 mm for tight coupling) in the coupled-line of which the accuracy is difficult to control in fabrication. And also the hand soldering of the resistor will also affected the measured results at higher frequencies. Figure 7-25 shows the predicted and measured $|S_{22}|$ and $|S_{32}|$. The measured $|S_{22}|$ at ports 2 and port 3 as well as $|S_{32}|$ between port 2 and port 3 are all lower than -10 dB over the realized UWB band.

Figure 7-23. Photograph of the Type-C UWB power divider. [78]

Figure 7-24. Predicted and measured results S_{21} - and S_{11} -magnitudes as well as group delay of the designed Type-C UWB power divider. [78]

Figure 7-25. Predicted and measured results S_{22} - and S_{32} -magnitudes of the designed Type-C UWB power divider. [78]

7.6 Conclusion

This Chapter has presented a class of UWB power dividers with good isolation between two output ports and sharp out-of-band roll-off skirt. A simple approach based on the equivalent circuit models under the even- and odd-mode excitations has been developed for the efficient analysis and design of three UWB power dividers. It was demonstrated that as two pairs of short-circuited stubs were introduced via an impedance transformer, three poles could be introduced in the UWB. As this transformer was replaced by the coupled line, two additional poles could be further excited. Furthermore, if the two pairs of short-circuited stubs are replaced by a shunt stepped-impedance open-stub, the rejection skirts could be further improved. After comparative study was executed for the three dividers, three prototype UWB dividers were designed and fabricated. Measured results confirmed the predicted power splitting performance with the 3-dB power transmission from one input to two outputs, good impedance matching at all the three ports, and good isolation between two outputs over the 3.1-to-10.6 GHz UWB band.

Chapter 8

Conclusion and Recommendations

8.1 Conclusion

UWB technology has a great potential in developing various modern transmission systems, however, the UWB bandwidth for indoor and outdoor communication systems are strictly regulated to the band from 3.1 GHz to 10.6 GHz by FCC part 15 regulations. To meet these requirements on the emission limit, it is highly demanded that UWB bandpass filters are developed to be able to cover the whole UWB passband with the fractional bandwidth of 110% at the center frequency of 6.85 GHz. So far, numerous UWB bandpass filters have been reported in the literatures with various design methodologies. In Chapter 2 of this thesis, some UWB bandpass filters were introduced with brief theoretical explanations and advantages/disadvantages were demonstrated.

The desired features for the bandpass filters used in UWB systems can be specified as, low insertion loss, high rejection, wide upper-stopband and small size. In Chapters 3-6, a number of UWB filters have been successfully developed in this thesis. They can be divided into four categories: lowpass embedded MMR UWB bandpass filters, EBG-

Conclusion

embedded UWB bandpass filter, short-circuit stub loaded MMR UWB bandpass filter, and notch-band UWB bandpass filter. They are all size miniaturized. The optimized filters are designed by Agilent Momentum software. The fabricated and measured results match the simulated results well.

Based on the ideas of aforementioned UWB bandpass filters, UWB power dividers were developed in Chapter 7 with good filtering performance or good out-of-band rejection. This was achieved by introducing short/open-circuit stub-loaded and coupled-line to the Wilkinson power divider. Three of the UWB power dividers with single isolation resistor were designed and fabricated to demonstrate and to confirm the predicted results.

The author has several main contributions in the UWB technology. Firstly, a synthesise method based on the Chebyshev function was developed for particular shape of MMR UWB filter. This method is able to extract the characteristic impedance of the MMR, and even and odd characteristic impedances of the coupled-line. Secondly, different MMR structures were investigated and embedded into the UWB filter. This not only provided a compact filter structure, but also creates a good upper-stopband. Thirdly, the design approach of three UWB power dividers was developed with closed-form formulae and design curves to provide useful design guidelines for UWB power dividers. Fourthly, all the proposed UWB filters and power dividers were designed on a single-layer PCB board, which was cost-effective and easy-to-fabricate. The filter size is very compact. The overall lengths of those proposed UWB filters are less than one guided wavelength. The smallest and the best performance UWB bandpass filter are around $0.5\lambda_g \times 0.25\lambda_g$ in Chapter 5.

8.2 Future Recommendations

- In this thesis, various multiple-mode resonators (MMRs) have been extensively characterized and discussed. There are still rooms for investigating different shapes of MMRs such that the performances of UWB bandpass filter can be further improved, aiming to sharpen out-of-band rejection skirts especially at lower stopband, produce multiple notched-bands in the UWB to avoid the affections to the existing wireless systems.
- All the proposed UWB filters in this thesis required very tight edge coupling degree between two coupled microstrip lines. Even though interdigital coupled-line has higher coupling than parallel coupled-line, its gap width is still very narrow according to the limited fabrication tolerance. For this purpose, coupled-line with backside aperture [17] and broadside coupled lines [25] can be employed to solve this problem and further applied for innovative design of these UWB filters with miniaturized size in advanced multilayered substrates, such as Low-Temperature Co-fired Ceramic (LTCC).
- Notch-band was initially created in a certain narrow band of the UWB passband by asymmetrically forming the coupled-line sections [34], [35], [42]. It can also be implemented in the MMR of the UWB filters by introducing a few quarter-wavelength open stubs into the MMR to produce multiple notched-bands as inquired in the wireless system.

Conclusion

- UWB power dividers with out-of-band improvement, proposed in this thesis, can be extended to develop UWB power dividers with 1 input port and N output ports. Furthermore, the coupled-section can be properly re-designed so as to construct an alternative UWB power divider with out-of-phase power division at its two output ports.

Publications:

Journal Papers:

- [1] **S. W. Wong** and L. Zhu, "UWB bandpass filter using shunt-stub embedded multiple-mode resonators," *Submitted Microwave and Optical Technology Letter* in February 2009.
- [2] **S. W. Wong** and L. Zhu, "Compact wideband bandpass filters using rhombus-shaped triple-mode resonator," *Accepted by Microwave and Optical Technology Letter* in December 2008.
- [3] **S. W. Wong** and L. Zhu, "Ultra-wideband power dividers with good isolation and improved sharp roll-off skirt," *Accepted by IET Microwaves, Antennas & Propagation, in January 2009.*
- [4] **S. W. Wong** and L. Zhu, "Quadruple-Mode UWB bandpass filter with improved out-of-band rejection," *IEEE Microwave Wireless Compon. Lett.*, vol. 19 , no. 2, pp. 152-154, March, 2009.
- [5] **S. W. Wong** and L. Zhu, "Ultra-wideband bandpass filters with improved out-of-band behavior via embedded electromagnetic-bandgap multimode resonators," *IET Microwaves, Antennas & Propagation*, vol. 2, no. 8, pp. 854-862, Dec. 2008.
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Conference papers:

- [1] **S. W. Wong** and L. Zhu, "Miniaturization of triple-mode ultra-wideband bandpass filters with good upper-stopband performance," in *Proceedings of 2008 International Microwave Workshop Series (IMWS'08)*, Chengdu China, December 2008.
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- [3] **S. W. Wong**, L. Zhu and S. Sun, "Ultra-wideband microstrip-line bandpass filters with good out-of-band performance using EBG-embedded multiple-mode resonator," in *Proceedings of 2007 Asia-Pacific Microwave Conference (APMC'07)*, Bangkok, December 2007.

Author's Publications

- [4] **S. W. Wong** and L. Zhu, "Ultra-wideband (UWB) microstrip bandpass filters with improved upper-stopband and miniaturized size," in *Proceedings of 2007 Asia-Pacific Microwave Conference (APMC'07)*, Bangkok, December 2007.
- [5] **S. W. Wong** and L. Zhu, "EBG-embedded UWB bandpass filter with widened upper-stopband," in *Proceedings of 2006 International Symposium on Antennas and Propagation*, November 1-4, 2006, Singapore, Paper code: a313_r298.
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Awards:

- [1] PhD Research Scholarship (2006-2010), Nanyang Technological University, Singapore.
- [2] 2007 IEEE Region 10 paper writing contest, NTU student branch, PG category, Second Price: "Ultra-Wideband Bandpass Filters with Improved Upper-Stopband Performance via Integration of Stub-Loaded Lowpass Structures."

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