Synthesis design of ultra-wideband (UWB) bandpass filters on planar transmission lines

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Synthesis Design of Ultra-Wideband (UWB) Bandpass Filters on Planar Transmission Lines

RUI LI

School of Electrical & Electronic Engineering

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

2009

DEDICATION

To

My supervisor

And

My grandparents, my parents and my husband

ACKNOWLEDGEMENTS

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

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SUMMARY

The objective of this PhD dissertation is to describe the design and implementation of high performance ultra-wideband (UWB) bandpass filters on planar transmission-line structures. With synthesis methodology, the insertion loss function of the proposed transmission-line model can be derived and analyzed to exhibit UWB bandpass or quasi-high-pass filtering behavior with Chebyshev equal-ripple responses in the main passband. Therefore, the element values involved in the networks can be calculated and the dimensions of the filter can be determined without tedious full-wave electromagnetic (EM) simulations. With geometrical variation of the initial filter, the overall circuitry size can be reduced and cross-coupling route can be introduced to enhance the filter selectivity. The synthesis approach offers the designers an alternative way to design UWB bandpass filters rapidly and efficiently.

After the literature survey of the conventional high-pass filter prototype with short-circuited shunt stubs, the dual network with series stubs cascading through uniform transmission lines is proposed and analyzed. By terminating the stubs with open/short-circuited ends, the novel high-/low-pass filter prototypes are established. By applying the synthesis design approach on the presented transmission-line models, the insertion loss functions of the prototypes are derived and regulated with the ideal generating functions that exhibit Chebyshev equal-ripple characteristics in the passband. Hence, the unknown element values in the transmission-line network can be obtained and tabulated as design references. The transmission-line models are then implemented on hybrid coplanar-waveguide and slotline structures or hybrid microstrip line and slotline structures to verify and validate the predicted results. Furthermore, the high order harmonic bands of

both the low-pass and high-pass filters can be effectively suppressed by replacing the uniform connecting lines between adjacent stubs with periodic structures.

Later on, the characteristics of stub-/slot-loaded multiple-mode resonators (MMR) are investigated. The electronic and magnetic planes on a MMR for different resonant modes are studied. By loading stubs/slots at those specific positions, the current path along the transmission-line resonator is rerouted. By adjusting the lengths of the loaded stubs/slots; the first few resonances can be moved close to each other and distributed evenly within the desired main passband. With the strong input/output coupling, the bandpass filter with wide/ultra-wide bandwidth can be constructed. Furthermore, with the help of a mode graph, the mode splitting behavior of multi-element resonator is studied and utilized to design multi-stage wide/ultra-wideband bandpass filters with improved rejection roll-offs and extended upper stopband.

Next, the slotline MMR is studied and used to construct UWB bandpass filters with microstrip-to-slotline transitions as distributive coupling structures at two ports. The resonance behavior of a uniform/non-uniform slotline resonator is discussed and made use of forming the ultra-wide passband with controllable bandwidth. The slotline resonator is then twisted to a W/bowl shape to miniaturize the overall filter size and to introduce an additional cross-coupling route between the input and output ports. Two transmission zeros are generated at the lower/upper passband edges to enhance the filter selectivity. To improve the upper stopband performance, low-pass structures are installed at two feed lines to suppress the high order harmonic bands.

The general transmission-line model with two sets of composite series and shunt stubs cascading through a multi-section transmission line is proposed and synthesized to

design UWB bandpass filters with designable in-band transmission poles. The insertion loss functions are derived directly from the presented network and regulated to exhibit Chebyshev equal-ripple responses in the passband. By equalizing the derived functions with the ideal generating functions, the element values involved in the transmission-line networks can be calculated and tabulated. The number of transmission poles within the passband can be increased by inserting more sections on the connecting line between two sets of stubs. Therefore, the prototype is capable to design UWB bandpass filters with sharp roll-offs. To verify and validate the predicted results, the proposed prototypes are implemented on hybrid microstrip line and slotline structures to realize the distinctive features of the transmission-line model. Furthermore, multi-stage transmission-line model with hybrid stubs is also presented and analyzed with synthesis approach. Compared with the traditional high-pass prototype, the proposed single/multi-stage networks provide higher-order transmission zeros and sharper rejection skirts at passband edges consequently.

ABBREVIATIONS

Balun	balanced to unbalanced transformers
EM	electromagnetic
EBG	electromagnetic bandgap
FCC	Federal Communications Commission
MICs	microwave integrated circuits
MMICs	monolithic microwave integrated circuits
MMR	multiple-mode resonator
PEW	perfect electric wall
PMW	perfect magnetic wall
RF	radio frequency
SIR	stepped-impedance resonator
UWB	ultra-wideband
WLAN	wireless local-area network

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CHAPTER 1

Introduction

1.1 Background

The descriptive term *microwave* is used to describe electromagnetic (EM) waves with frequency ranging from 300 MHz to 30 GHz, with the corresponding electrical wavelength range of 1 m to 1 cm in free space. Over the past one hundred years, microwave has become a ubiquitous technology that is used not only for radar and communication systems, but also for the characterization and analysis of materials, industrial drying, environmental remote sensing, medical diagnosis and treatment and so on [1]-[6].

Since 1950s, the emergence of microwave integrated circuits (MICs), especially monolithic microwave integrated circuits (MMICs), has been playing a very important role in the advanced radio frequency (RF) & microwave technologies [7]-[9]. Over the

Fig. 1.1. Cross-sectional views of planar transmission lines for MICs. (a) Stripline. (b) Microstrip line. (c) Conductor-backed coplanar-waveguide. (d) Coplanar-waveguide. (e) Slotline. (f) Coplanar stripline.

past 50 years, MICs and MMICs have experienced a tremendous growth. Circuits become more compact, lighter in weight, highly integrated, lower cost and have found extensive applications in radar, electronic warfare, and the commercial field. The transmission structure in MICs should be “planar” in configuration [10]-[12]. A planar configuration implies that the characteristics of the element can be determined by the dimensions in a single plane. The most common planar transmission structures are stripline, microstrip line, coplanar waveguide, slotline and coplanar stripline, etc [13]-[18]. Cross-sectional views of these transmission lines are shown in Fig. 1.1. These planar structures act as the backbone of MICs and offer an efficient and flexible way in the design of MICs.

Microstrip line is one of the most popular types of planar transmission lines, primarily because it can be fabricated by photolithographic processes and be easily integrated with other passive and active microwave devices. The filters implemented on printed microstrip line structures have the advantages of compact size, simple structure as well as low cost. In this part, some conventional microstrip line low-pass/bandpass filters will be reviewed.

Stepped-impedance low-pass filter is a direct implementation of traditional L-C low-pass prototype on microstrip line structure [19]. It consists of alternating high- and low-impedance transmission lines, which are much shorter than the associated guided-wavelength. So it is considered as semi-lumped element low-pass filter. As compared with L-C ladder type of low-pass prototype, the high-impedance lines act as series inductors and the low-impedance lines act as shunt capacitors. An alternative realization of a shunt capacitor is to use an open-circuited stub with a physical length that is shorter than a quarter of guided wavelength at cutoff frequency. The performances of these two filters are approximate only in the vicinity of the cutoff frequency but not in the wide frequency range.

Parallel-coupled microstrip bandpass filters use half-wavelength line resonators edge-coupled with each other [19]. They are positioned so that adjacent resonators are parallel-coupled along half of their length. Hairpin line bandpass filter is a compact structure and it can be considered to be obtained by folding the resonators of parallel-coupled bandpass filter into a U shape [20]-[23]. Conventional hairpin line filters were firstly introduced by Cristal and Frankel [20] in the early 1970's. They were designed to meet the system demand which requires smaller and lighter filters. In the later 1980's,

further miniaturized hairpin resonator filters were reported for the application of receiver based on the front-end MICs [21]. In [22], a novel class of cross-coupled bandpass filter with modified hairpin resonators was proposed. The cross-coupled filter exhibits expected ripples at the passband and the stopband so as to improve frequency selectivity and reduce in-band loss. Later on, the cross-coupled hairpin resonators with compact size were constituted to design dual-band bandpass filter with quasi-elliptic responses [23].

Interdigital bandpass filter was initially presented in [24], which consists of an array of quarter-wavelength resonators at the midband frequency. The resonators are edge-coupled and each of them has one short-circuited end and one open-circuited end. In such an orientation, the type of filter is extremely compact. However, it requires the use of grounded microstrip resonators which usually needs to be achieved with via holes. Since the resonators are quarter-wavelength with one short-circuited end, the second passband of the filter is at three times far away from the central band, thus accomplishes wide upper stopband. Later on, with the under-coupled resonators [25] and stepped-impedance resonators [26], the second harmonic band is suppressed or shifted beyond three times of the fundamental passband as expected.

The concept of a microstrip dual-mode bandpass filter was firstly presented in 1972 [27], extensive works have been done so far to construct various types of bandpass filters by exciting the two degenerate modes of ring resonators [28]-[31] and patch resonators [32] –[35]. By properly installing some perturbation elements such as cuts or stubs in either of diagonal planes, these two degenerate modes can be simultaneously excited and they can be employed as two resonators in filter design. Hence, the overall size of dual-mode resonator filters can be dramatically reduced by about 50% or more, thus making

up a compact filter configuration. Over the past several years, research works have been extensively made to improve the performances of ring resonator based dual-mode filters, such as harmonic suppression and size miniaturization [28]–[31]. In [28], the stepped-impedance ring resonator and spur-line loaded feed lines are utilized together to put off and suppress the lowest spurious resonant mode, respectively. Later on, four open-ended stubs [29] and four capacitive small patches [30] are etched in the four corners of a rectangular ring resonator to simultaneously achieve size reduction and harmonic suppression with the use of attractive slow-wave properties. In [31], a called periodic stepped-impedance ring resonator is proposed based on the similar concept to explore a super small and wide stopband dual-mode ring filter. As an alternative resonator, a square patch with crossed slots is initially proposed in [32] to make up a new type of dual-mode filters with high power-handling efficiency [33]. With the proper choice of two perpendicular crossed slots in length, this dual-mode filter can achieve simultaneous size and loss reduction. These two aspects of performance get the further improvement by forming an inductively loaded cross-slotted patch resonator [34]. Recently, the work in [35] pays its attention on the existence of narrow upper stopband in the initial cross-slotted resonator and proposes a modified resonator configuration by rotating the cross slots in parallel to the two perpendicular feed lines so as to solve this problem.

In the modern days, the rapid growth in fabrication technologies and the successful commercial deployment of wireless communications have significantly affected our daily lives. The arrival of the third- and fourth-generation radio systems and the replacement of wired connections with Wi-Fi and Bluetooth enable people to access to a wide range of information from anywhere and at any time. As the demand for higher capacity and faster

Introduction 6

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service increases, new technology has to be found in the crowded and scarce RF spectrum. Since most of the interested frequency bands have been occupied by many specific usages, it becomes stricter with the introduction of new technology without interfering with the existing applications.

Ultra-wideband (UWB) technology offers a promising solution to the RF spectrum shortage by allowing new services to coexist with current radio systems with minimal or no interference [36]-[42]. UWB wireless communication is not a new term and it was studied in the past for different purposes. It was initially investigated in the late 1960s by Henning Harmuth of Catholic University of America and Gerald Ross and K. W. Robins of Sperry Rand Corporation [36]. From 1960's to the 1990's, this UWB technology was restricted to military and defense applications under various classified programs, such as highly secure communications [37]. The recent development of high speed microprocessors and fast switching techniques make this UWB technology possible for commercial short-range communications [40]-[42].

UWB systems transmit extremely short pulses 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 current radio services. Besides the ability of sharing the frequency spectrum, UWB signals have additional advantages such as,

- Large channel capacity makes UWB system a perfect candidate for short-range, high-data-rate wireless applications.

- Low power density that is usually below environment noise ensures communication security.
- The short-duration pulse lowers the sensitivity to multi-path effect.
- Carrierless UWB transmission simplifies the transceiver architecture and thus reduces the cost.

In February 2002, Federal Communications Commission (FCC) authorized the unlicensed commercial deployment of UWB technology under a strict power control in 7500 MHz spectrum [43]. 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

(a)

(b)

Fig. 1.2 UWB emission limits. (a) Indoor systems. (b) Outdoor systems.

communication is currently considered as one of the most promising areas [41].

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 and above 10.6 GHz. The emission limits for indoor and outdoor UWB devices are shown in Fig. 1.2 (a) and (b), respectively. 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 FCC also dictates that no fixed infrastructure can be used for UWB communications in outdoor environments. Therefore, outdoor UWB communications are restricted to handheld devices that can send information only to their associated receivers.

In UWB communication systems, bandpass filter plays an important role, which should exhibit a wide bandwidth with low insertion loss over the whole band. In order to meet the FCC limit, good selectivity at both lower and higher frequency ends and flat group delay response over the entire band are required. Great interest has been aroused toward exploring UWB bandpass filters with various desired features in both academic and industrial areas. Some typical types of UWB bandpass filters proposed in recent years will be introduced in the following Section.

1.2 A Brief Literature Review of UWB Bandpass Filter

The conventional well-known high-pass prototype with short-circuited shunt stubs [44] is commonly adopted in designing bandpass filters with ultra-wide bandwidth [45]-[50]. Following that, an ultra-wideband bandpass filter was initially presented in [45] and then was systematically optimized by cascading such high-pass sections with an active isolation and a low-pass section [46]. However, the resultant filter occupied a large circuitry area. Hence, extensive work has been carried out towards miniaturizing the overall filter size and improving the electrical performance at the same time. In [47], with the electromagnetic bandgap structure formed on the connecting line between two adjacent stubs, the upper stopband was effectively extended and the total circuitry size was miniaturized. Later on, a cross-coupling route was purposely introduced between the input and output ports in [48] and [49], resulting in enhanced filter selectivity and improving the group delay performance. By adding two quart-wavelength transmission-line sections at two ports and replacing the connecting line by a low-pass structure, a five-pole UWB filter with widened upper stopband was constructed in [50]. This

prototype has been recently employed to design a UWB filter on a multilayer liquid-crystal-polymer substrate [51]-[53]. However, in constructing these filters on microstrip line topology, via holes are always required to realize the short-ended terminals for all shunt stubs.

Another major class of UWB bandpass filters based on the so called multiple-mode resonator (MMR) theory becomes quite popular due to its simple structure and easy design procedure [54]-[62]. The stepped version of the MMR is also commonly referred to as stepped-impedance resonator (SIR). In this dissertation, the term MMR is used to indicate a resonator of straight, bent or stepped shape. The concept of using multiple resonant frequencies to construct the ultra-wide passband was firstly introduced in [57] and then implemented with different transmission-line structures [58]-[62], [67]-[71]. The basic idea of this kind of UWB filter is to construct the ultra-wide passband with the first few resonant frequencies. The locations of those resonances are controlled by the geometrical properties of the MMR. One drawback of this kind of bandpass filters is that it has relatively narrow upper stopband due to the higher order harmonic bands. Therefore, the electromagnetic bandgap structures embedded resonator was employed to suppress the harmonic band successfully and reduced the overall filter size at the same time in [69]-[71]. On the other hand, low-pass or bandstop structures were used to improve the out-of-band performance in [62]-[74]. Although the UWB applications do not affect the existing radio systems, the exclusive use of the spectrum such as wireless local-area network (WLAN) may interfere with the UWB range defined by FCC. To circumvent this problematic issue, a single [75]-[80] or multiple [81] narrow-band notched UWB filters were developed.

There were also other types of UWB bandpass filters designed with different approaches [82]-[85]. The resonant modes of stepped-impedance resonator and split modes graph of cascading resonators were studied in detail in [90]-[91]. The resonant peaks of the multi-stage structure were matched with the nominal Chebyshev poles. The traditional edge-coupled line structure in this type of microstrip line filter may not be strong enough and it is limited by the tolerance in fabrication process. Therefore, the broadside coupling structures become one solution for this problem [92]-[93].

1.3 Motivation

UWB technique is fundamentally different from all other communication techniques because it employs extremely narrow RF pulses to communicate between transmitters and receivers. The short-duration pulses generate a very wide bandwidth and give several advantages, such as large throughput, robustness to jamming, and coexistence with current radio services. After the FCC released the unlicensed use of UWB devices in the frequency band 3.1 – 10.6 GHz, great interest in UWB communications has been aroused from both academic and industrial areas [37]-[42] and numerous UWB devices have been designed and implemented [96]-[101]. As an essential component in all wireless communication systems, high performance microwave filters are highly demanded in regulating the frequency spectrums and suppressing the harmonics generated by the non-linear devices. As shown in Fig. 1.2 (a) and (b), the FCC spectrum masks for the UWB indoor and outdoor applications are quite strict. Therefore, microwave bandpass filters with ultra-wide bandwidth, sharp rejection skirts at both passband edges as well as wide upper stopband are required to regulate the frequency spectrum to fulfill the given masks.

A variety of UWB bandpass filters with high selectivity, flat group delay, extended upper stopband and compact size have been reportedly developed as summarized above. Nevertheless, these filters are usually designed relying on the cut-and-try approach where the overall filter layout with electrically large size is simulated again and again using the commercial software towards the specified ultra-wide passband. It is a time and effort consuming process, which lengths the design cycles a lot. As a result, there is a demand to develop a systematic synthesis approach such that these bandpass filters with a fractional bandwidth of 110% can be efficiently designed. As the conventional low-pass filter prototype, all the element values of the prototype can be derived and tabulated with different design specifications. The designer, who is asked to implement an UWB bandpass filter with required bandwidth and in-band return loss, can simply read the design parameters from the table and then implement it on transmission-line structures. With optimization fine tuning, a high performance filter fulfilling the design requirement can be constructed without tedious full-wave EM simulation. On the other hand, as the common drawback of all the synthesis methods, only simple and straight forward structures could be designed by synthesis. However, in modern wireless communication systems, other features such as compact size, extendable upper stopband and high selectivity are also desired. Therefore, the initial dimensions calculated from the synthesis procedure can be used a perfect starting point or basis for adding more features. Geometrical variation is always needed to achieve the target and improve the filter's performance.

1.4 Objectives

With the motivation mentioned above, the goal of this dissertation is to develop a synthesis methodology to design microwave bandpass filters with ultra-wide bandwidth and to implement the proposed transmission-line models on planar transmission-line structures. Therefore, high performance UWB bandpass filters that regulate the spectrum to meet the FCC requirements can be designed more rapidly and efficiently without cut-and-try full-wave EM simulation. To further improve the performance of the UWB filters, geometrical variation of the basic synthesized filters might be needed. As mentioned in the literature review of the UWB bandpass filter in Section 1.2, UWB filters can be designed using the conventional high-pass filter prototype, which is analyzed by a synthesis approach. This dissertation will start with this synthesis procedure and make use of it to analyze the proposed transmission-line models which exhibit UWB filtering responses.

1.4.1 Low-/high-pass prototype with series stubs

- To analyze the well-known conventional high-pass filter prototype with shunt short-circuited stubs cascading through uniform transmission lines and to study the characteristics of the insertion loss function that exhibits Chebyshev equal-ripple responses.
- To propose a “dual network” of the traditional high-pass prototype that consists of series stubs terminated with short-/open-circuited ends and uniform transmission lines that connect the adjacent stubs.

- To analyze the proposed transmission-line model with synthesis design mythology and to regulate the insertion loss function with ideal targeted generating function with Chebyshev equal-ripple characteristics.
- To implement the presented transmission-line network on planar transmission-line structures and to measure the fabricated prototype filters in order to validate and confirm the theoretical predictions.

1.4.2 UWB bandpass filters with improved performance

- To characterize the equivalent circuit model of series open-circuited stubs and to study the connection between the high-pass prototype with series open-ended stubs and the UWB filter with MMR.
- To analyze the characteristics of the stub-/slot-loaded MMR based on the multiple-mode resonator theory.
- To design the UWB bandpass filter with the first few resonances of the stub-/slot-loaded MMR with controllable bandwidth.
- To miniaturize the size of the proposed UWB bandpass filter.
- To enhance the filter selectivity and to extend the upper stopband of the UWB bandpass filter.

1.4.3 UWB bandpass filter with slotline MMR

- To extend the design theory of UWB bandpass filter with microstrip MMR to slotline resonator and to study the characteristics of slotline MMR.

- To study the characteristics of microstrip-to-slotline transition as a distributed coupling structure.
- To construct and implement UWB bandpass filter with slotline MMR and microstrip-to-slotline transitions.
- To further improve the filter performance and make it fulfill the FCC regulations.

1.4.4 High-pass filter prototype with hybrid series and shunt stubs

- To propose a general single stage high-pass prototype with composed series and shunt stubs connecting through a multi-section non-uniform transmission line.
- To analyze the presented network with synthesis methodology. To derive the insertion loss function of the proposed high-pass prototype and to characterize its frequency responses.
- To regulate the insertion loss function with the ideal functions exhibiting Chebyshev equal-ripple characteristics and to determine the design parameters of the proposed transmission-line model.
- To propose a multi-stage high-pass prototype with hybrid series and shunt stubs and connecting lines and to analyze it with synthesis approach.
- To implement the proposed networks on planar transmission-line structures and to measure the fabricated filter circuits to verify and validate the theoretical predictions.

1.5 Major Contributions

The research outcome achieved in this PhD. dissertation represents the innovative design of microwave ultra-wideband bandpass filters using synthesis design approach.

The major contributions of this dissertation are listed as follows:

- ✓ Proposed transmission-line models for low-/high-pass filters with Chebyshev equal-ripple responses in the passband. The presented networks consist of series open-/short-circuited stubs cascading through uniform transmission-line sections. The insertion loss functions of the transmission-line models are derived and regulated with ideal Chebyshev generating functions. The element values involved in the networks are calculated and tabulated for other designers' reference. The low-/high-pass prototype filters are implemented on hybrid coplanar-waveguide and slotline structures to validate the predicted results. The proposed transmission-line models are a dual of the traditional low-/high-pass prototypes with shunt short-/open-circuited stubs, and provide an alternative way to design and implement UWB bandpass filters. The element values obtained from synthesis approach make the filter design rapid and efficient without going through the tedious full-wave EM simulation. The dimensions of the filter can be easily determined and only fine tunings are needed to get the optimized performance.
- ✓ Constructed the novel stub-loaded UWB bandpass filter with miniaturized circuitry size and extended upper stopband. The resonant frequencies of a transmission-line MMR can be influenced by loading stubs along the resonator at some specified positions. The stub-loaded MMR is proposed and its first four

resonances are moved close to each other by adjusting the lengths of the loaded stubs. With strong interdigital coupling applied at input/output ports, an UWB bandpass filter is constructed. Based on similar idea, slot-loaded MMR is presented to design wide/ultra-wideband bandpass filters. Mode-graph is studied and utilized to design bandpass filters with multi-stage slot-loaded MMRs. The filter selectivity is effectively enhanced and the upper stopband is widened.

- ✓ Characterized the resonance behavior of uniform/non-uniform slotline MMR. Proposed UWB bandpass filters with the first three resonant frequencies of slotline MMR and microstrip-to-slotline transitions as frequency distributed external coupling structure. By twisting the slotline MMR to W-/bowl-shape, cross coupling routes between the input and output ports are successfully introduced. Two transmission zeros at both passband edges are created to sharpen the rejection skirts. The upper stopband performance of the UWB filters is improved by loading low-pass structures at two feedlines. Therefore, novel UWB bandpass filters with compact size, enhanced selectivity and wide upper stopband are established.
- ✓ Proposed a general single stage transmission-line mode of UWB bandpass filter with Chebyshev equal-ripple response in the passband. The presented network is composed of two sets of hybrid series and shunt stubs connecting through a multi-section transmission line. The number of transmission poles in the passband can be controlled by the number of transmission-line sections between the two sets of stubs. By applying the synthesis methodology on the proposed transmission-line model, the insertion loss function is derived and regulated

with the targeted generating function. Hence, the element values in the networks can be calculated and tabulated as future design guidelines. Multiple-stage transmission-line model is also proposed and analyzed with similar method. Both the single and multiple stage UWB bandpass filters offer high-order transmission zeros at two sides of the passband as compared with the traditional high-pass filter prototype. Therefore, shaper roll-offs can be achieved without increase the filter order. Furthermore, the synthesis designed UWB filter provides an alternative view of the MMR UWB filter and gives the designers a better understanding of the filter theory.

1.6 Organization of the Dissertation

The PhD dissertation is organized as follows. Chapter 2 briefly introduces the conventional high-pass prototype and proposes its dual network. The transmission-line model is analyzed with synthesis methodology and the element values are calculated and tabulated. The networks are implemented on hybrid coplanar waveguide and slotline structures to experimentally confirm and verify the predicted results.

Chapter 3 investigates the characteristics of stub-/slot-loaded MMRs. The resonance behaviors with respect to the stub/slot loading positions are discussed and the UWB bandpass filters making use of these properties are proposed, designed and realized. Design examples are given and confirmed by full-wave electromagnetic simulation and experimental measurement.

Chapter 4 presents a novel UWB bandpass filter with slotline resonator and microstrip-to-slotline transitions. The characteristics of the uniform/non-uniform slotline

resonator are investigated and used to design bandpass filters with controllable ultra-wide bandwidth. With the geometrical variation based on the basic filter structure, cross-coupling route is introduced to enhance the filter selectivity. The designed W/bowl-shape UWB bandpass filters demonstrate good filtering behavior with sharp roll-off and wide upper stopband.

Chapter 5 thoroughly discusses the synthesis design procedure of the proposed transmission-line models for UWB bandpass filters to exhibit Chebyshev equal-ripple responses within the passband. With the synthesis analysis approach, the element values involved in the networks can be determined. The presented prototypes are then implemented on hybrid transmission-line structures to verify and validate the theoretical predictions.

Finally, conclusions and recommendations are made in Chapter 6.

CHAPTER 2

Synthesis Design of Microwave Low-/High-Pass Filters with Series Stubs

2.1 Introduction

A microwave filter is a two-port component used to control the frequency response at a certain point in a microwave system by providing transmission at frequencies within the passband of the filter and attenuation in the stopband of the filter [5]-[6], [19]. In general, microwave filters are categorized into four types: low-pass, high-pass, bandpass and bandstop, and their applications can be found in all kinds of microwave communication systems. There are mainly two ways to design a microwave filter, *image parameter method* and *insertion loss method* [5]. The former was developed in the late 1930s and was useful for low-frequency filters in radio and telephony. While the later is

considered as a more modern procedure emerged in early 1950s, and it is a base of most sophisticated computer-aided microwave filter design packages used today.

The *insertion loss method* starts with low-pass filter prototypes consisting series and shunt elements that are normalized in terms of impedance and frequency. After determination of all the element values in the networks, the low-pass prototypes can be scaled to the desired frequency and impedance. Therefore, all the other types of filters such as high-pass, bandpass and bandstop filters operating at arbitrary frequency and port impedance can be obtained. The design process is illustrated in Fig. 2.1.

Fig. 2.1 The process of filter design by the insertion loss method.

The lump-element filter design theory only works well at lower frequencies; however, this theory becomes inapplicable and inaccurate when the frequency becomes higher. Lumped-element circuit analysis assumes that the physical dimensions of a network are much smaller than the electrical wavelength. As the frequency increases, the wavelength becomes much shorter. Hence, the distances between filter components are not negligible. Besides that, the lumped elements such as inductors and capacitors are generally available only for a limited range of values and are difficult to implement at microwave frequencies. To bridge the gap, Richard's transformation [102] was developed to convert the lumped-element network to its relevant transmission-line network, while Kuroda's identities were applied to rearrange all the elements involved in the filter by

adding the redundant transmission-line sections. Therefore, the lumped-element network theory has been successfully extended to microwave networks.

In the middle 1960s, a lot of work had been done to develop an alternative direct synthesis method based on the realistic distributed-element and/or transmission-line networks [103]-[111]. In [105], Riblet introduced a simple explicit transfer function in terms of Chebyshev polynomials based on a general rational function given by Sharpe [106]. As such, a class of N^{th} -order transmission line filters could be constituted by cascading N -section shunt-resonant elements through $(N-1)$ -section quarter-wavelength transmission lines and its frequency response could be easily specified as an equal-ripple function. On the other hand, an alternative approach was presented by Carlin and Kohler in [111] to constitute a bandpass filter by cascading the mixed shunt and series stubs via uniform line. A generic insertion loss transfer function was developed in [109] and [111] to carry out the synthesis design of a branch-guide directional coupler [44] and bandpass filter [108]. In [108], the general prototype network was constructed by cascading shunt stubs of electrical length θ at a specified frequency via uniform lines of length 2θ at the same frequency. This transmission-line based filter network could be mixed by shunt lumped elements and synthesized to derive low-/high-pass responses with quasi-Chebyshev characteristics.

2.2 Synthesis Design of Low-/High-Pass Filters with Shunt Stubs

The low-/high-pass prototypes presented in [108] consist of n shunt open-/short-circuited stubs cascading through $(n - 1)$ uniform stubs. The electrical lengths of all the

Fig. 2.2. Transmission-line models of the conventional filter prototypes based on shunt stubs. (a) Low-pass prototype with open-circuited shunt stubs. (b) High-pass prototype with short-circuited shunt stubs.

shunt stubs and connecting lines are fixed as θ_c and $2\theta_c$ at specified cutoff frequency f_c . The characteristic impedances of all the involved transmission line structures are normalized with respect to the port impedance, the normalized impedances for the stubs and the connecting lines are indicated by a_i ($i = 1, 2, \dots, n$) and b_i ($i = 1, 2, \dots, (n-1)$), and the left and right ports are terminated by two unit normalized impedances, i.e., $z_0 = 1$. The transmission-line models for these two prototypes are depicted in Fig. 2.2 (a) and (b). Richard's variable can be defined as below for a lossless network,

$$t = j \tan \theta \quad (2.1)$$

For the low-pass prototype with shunt open-circuited stubs shown in Fig. 2.2(a), the transfer matrix of an open-circuited series stub with normalized characteristic impedance a_i is expressed as

$$\begin{bmatrix} 1 & 0 \\ t/a_i & 1 \end{bmatrix} \quad (2.2)$$

and the transfer matrix of a uniform transmission line with normalized characteristic impedance b_i is given by

$$\frac{1}{1-t^2} \begin{bmatrix} 1+t^2 & 2b_i t \\ \frac{2}{b_i} t & 1+t^2 \end{bmatrix} \quad (2.3)$$

Therefore, the overall normalized $ABCD$ matrix for the low-pass prototype shown in Fig. 2.2(a) can be derived by multiplying the individual $ABCD$ matrix of each transmission-line section in sequence,

$$\frac{1}{(1-t^2)^{n-1}} \begin{bmatrix} A_{n-1}(t^2) & tB_{n-2}(t^2) \\ tC_{n-1}(t^2) & D_{n-1}(t^2) \end{bmatrix} \quad (2.4)$$

where the subscripts indicate the degree of polynomials $A(t^2)$, ..., $D(t^2)$ with the even-odd variable t^2 . The values of the reflection coefficient Γ and transmission coefficient T for a network are given in terms of the $ABCD$ matrix parameters by the well-known formulas [112],

$$\Gamma = \frac{(A-D) + (B-C)}{A+B+C+D} \quad (2.5a)$$

$$T = \frac{2}{A+B+C+D} \quad (2.5b)$$

For a lossless network, the relationship of these two coefficients is as follows,

$$|\Gamma|^2 + |T|^2 = 1 \quad (2.6)$$

Therefore, by dividing both sides of (2.6) with $|T|^2$, the insertion loss function P_L can be obtained as,

$$P_L = 1 + \left| \frac{\Gamma}{T} \right|^2 \quad (2.7)$$

By substituting (2.5) into (2.7), the insertion loss function is expressed in terms of transfer matrix of the entire network,

$$P_L = 1 + \left| \frac{(A - D) + (B - C)}{2} \right|^2 \quad (2.8)$$

For a symmetrical network with reference to two external ports as shown in Fig. 2.2 (a) and (b), the two entries A and D in the transfer matrix are equal to each other. Hence, the insertion loss function in (2.8) can be further simplified as

$$P_L = 1 + |F|^2 \quad (2.9)$$

where $F = (B - C)/2$ and it is defined as a characteristic function. By substituting B and C from (2.4) into function F ,

$$F = \frac{t(B_{n-2}(t^2) - C_{n-1}(t^2))}{2(1 - t^2)^{n-1}} = \frac{t(P_{n-1}(t^2))}{(1 - t^2)^{n-1}} \quad (2.10)$$

where P_{n-1} is a polynomial of degree $(n - 1)$. By replacing the Richard variable in (2.10) with (2.1), the insertion loss function can be derived as,

$$P_L = 1 + \left| \frac{P_{2n-1}(\sin \theta)}{\cos \theta} \right|^2 \quad (2.11)$$

In order to achieve the Chebyshev equal-ripple response in the band $-x_c \leq x \leq x_c$ where $x = \sin \theta$, $x_c = \sin \theta_c$, P_L should be written in such form [105], [108],

$$P_L = 1 + h^2 \left\{ \frac{(1 + \sqrt{1 - x_c^2})T_{2n-1}\left(\frac{x}{x_c}\right) - (1 - \sqrt{1 - x_c^2})T_{2n-3}\left(\frac{x}{x_c}\right)}{2\sqrt{1 - x^2}} \right\}^2 \quad (2.12)$$

where $T_n(x)$ is the Chebyshev polynomial function of the first kind of degree n and h is the specified equal-ripple constant in the concerned frequency band. It relates with ripple level L_A in such a way that,

$$L_A = 10 \log_{10} (1 + h^2) \quad (2.13)$$

By equalizing (2.11) with (2.12), the element values for the shunt open-circuited stubs and the connecting lines can be calculated once the specified cutoff phase θ_c and ripple constant h are set. For the high-pass prototype with short-circuited shunt stubs shown in Fig. 2.2(b), the derived insertion loss function and the targeted one can be simply obtained by replacing θ in (2.11) with $(90^\circ - \theta)$ and to define $x = \cos\theta$, $x_c = \cos\theta_c$ in (2.12). As such, all the relevant element values for the high-pass prototype can be determined and partially tabulated in [113] for $n = 2, 3, 4, 5, 6$ and $\theta_c = 25^\circ, 30^\circ, 35^\circ$.

2.3 Synthesis Design of Low-/High-Pass Filters with Series Stubs

2.3.1 General transmission-line models

As a “dual network” of the conventional low-/high-pass prototype filters with shunt stubs, the proposed transmission-line models are composed of n series short-/open-stubs with electrical length θ_c at specified cutoff frequency f_c , cascading through $(n - 1)$ uniform transmission lines with length $2\theta_c$ as depicted in Fig. 2.3. Again, the normalized characteristic impedances of the series stubs and connecting lines are indicated as a_i ($i = 1, 2, \dots, n$) and b_i ($i = 1, 2, \dots, (n - 1)$) as those in Fig. 2.2. The transmission-line networks depicted in Fig. 2.3 (a) and (b) are the low-pass and high-pass prototypes, respectively. Following the synthesis procedure that is presented in Section 2.2, the overall transfer matrix of the whole network in Fig. 2.3(a) can be derived easily as,

Fig 2.3. Transmission-line models of the proposed filter prototypes based on series stubs. (a) Low-pass prototype with short-circuited series stubs. (b) High-pass prototype with open-circuited series stubs.

$$\frac{1}{(1-t^2)^{n-1}} \begin{bmatrix} A_{n-1}(t^2) & tB_{n-1}(t^2) \\ tC_{n-2}(t^2) & D_{n-1}(t^2) \end{bmatrix} \quad (2.14)$$

By comparing (2.14) with (2.4) for shunt stubs, it is noticed that two entries A and D keep the same, while the two entries B and C are exchanged with each other. Therefore, the characteristic function of the proposed prototype filters with series stubs has opposite sign or is out of phase with that with shunt stubs in (2.10), and it is shown as follows,

$$F = \frac{t(B_{n-1}(t^2) - C_{n-2}(t^2))}{2(1-t^2)^{n-1}} \quad (2.15)$$

However, the insertion loss function P_L only relates function F with its magnitude, it implies that the insertion loss function for these two prototypes shown in Fig. 2.2(a) and Fig. 2.3(a) are the same. Therefore, the proposed transmission-line models provide an alternative choice of designing low-/high-pass filters. To calculate the element values in

the network shown in Fig. 2.3(a), the targeted insertion loss function should have the format shown below,

$$P_L = 1 + h^2 \left\{ -\frac{(1 + \sqrt{1 - x_c^2})T_{2n-1}\left(\frac{x}{x_c}\right) - (1 - \sqrt{1 - x_c^2})T_{2n-3}\left(\frac{x}{x_c}\right)}{2\sqrt{1 - x^2}} \right\}^2 \quad (2.16)$$

where the variable x and x_c have the same definition as those in (2.12). The only difference between (2.16) and (2.12) is the additional minus sign in the term within the bracket. Therefore, by equalizing the insertion loss function derived from the transmission-line model in Fig. 2.3(a) with (2.16), the element values a_i and b_i can be calculated. Appendix I and II tabulate the element values of Chebyshev low-pass filter prototypes with $n = 2$ and 3 under the ripple levels L_A of 0.1 and 1.0 dB, respectively. For the high-pass prototype with open-circuited series stubs shown in Fig. 2.3(b), if θ in derived insertion loss function for low-pass prototype is replaced by $(90^\circ - \theta)$ and redefine $x = \cos\theta$, $x_c = \cos\theta_c$ in (2.16), the derived and targeted insertion loss functions are obtained. All the relevant element values for the high-pass prototype can be simply generated by replacing θ_c with $(90^\circ - \theta_c)$ in Appendix I and II.

2.3.2 Design example of low-pass prototype

For the low-pass prototype shown in Fig. 2.3(a) with $n = 2$, the normalized impedance for two series stubs become exactly the same, i.e., $a_1 = a_2$, due to the even-symmetrical geometry. The overall $ABCD$ matrix of the transmission-line model is as following,

Fig. 2.4. Normalized impedances (a_1 and b_1) as a function of the cutoff electrical length (θ_c) for a low-pass prototype with $n = 2$ and $L_A = 0.1$ dB. k_1, k_2 : Defined in (2.20).

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{n=2} = \begin{bmatrix} \cos 2\theta - \frac{a_1}{b_1} \tan \theta \sin 2\theta & j2a_1 \cos 2\theta \tan \theta - j \frac{a_1^2}{b_1} \tan^2 \theta \sin 2\theta + jb_1 \sin 2\theta \\ j \frac{1}{b_1} \sin 2\theta & \cos 2\theta - \frac{a_1}{b_1} \tan \theta \sin 2\theta \end{bmatrix} \quad (2.17)$$

In this case, the function F can be simplified to that in (2.18) with an odd-polynomial of $\sin \theta$ in its numerator. To realize the specified Chebyshev equal-ripple response, the magnitude of F can be obtained as (2.19) based on the formula in (2.16).

$$F = j \frac{(-4a_1 - 2b_1 + \frac{2}{b_1} - \frac{2a_1^2}{b_1}) \sin^3 \theta + (2a_1 + 2b_1 - \frac{2}{b_1}) \sin \theta}{2 \cos \theta} \quad (2.18)$$

$$|F| = \left| -h \frac{(1 + \sqrt{1 - \sin^2 \theta_c}) T_3\left(\frac{\sin \theta}{\sin \theta_c}\right) - (1 - \sqrt{1 - \sin^2 \theta_c}) T_1\left(\frac{\sin \theta}{\sin \theta_c}\right)}{2 \cos \theta} \right| \quad (2.19)$$

By equalizing the coefficients of $\sin^3 \theta$ and $\sin \theta$ in (2.18) and (2.19), a coupled set of equations can be derived as below,

$$2a_1 + b_1 - \frac{1}{b_1} + \frac{a_1^2}{b_1} = \frac{2h(1 + \sqrt{1 - \sin^2 \theta_c})}{\sin^3 \theta_c} = k_1 \quad (2.20a)$$

$$-a_1 - b_1 + \frac{1}{b_1} = -\frac{h(2 + \sqrt{1 - \sin^2 \theta_c})}{\sin \theta_c} = k_2 \quad (2.20b)$$

By summing (2.20a) and (2.20b), b_1 is expressed as a function of a_1 as well as two additional coefficients k_1 and k_2 that are defined in (2.20a) and (2.20b).

$$b_1 = \frac{a_1^2}{k_1 + k_2 - a_1} \quad (2.21)$$

In (2.21), the normalized impedance b_1 (for the connecting line) must be positive, *i.e.*, $b_1 > 0$, so the normalized impedance a_1 (for the series stubs) should satisfy

$$a_1 < k_1 + k_2 \quad (2.22)$$

Under $L_A = 0.1$ dB, three curves of a_1 , b_1 and $(k_1 + k_2)$ against θ_c are plotted in Fig. 2.4. As θ_c increases, a_1 gradually goes down and approaches its upper limit $(k_1 + k_2)$ as restricted in (2.22). If θ_c exceeds 70° , a_1 is larger than its limit such that (2.22) becomes invalid and the normalized impedance b_1 becomes negative. Therefore, it is a non-physical case to design a low-pass filter with the cutoff electrical length $\theta_c > 70^\circ$. Similarly, it can be figured out that the high-pass filters with $\theta_c < 20^\circ$ are also not realizable in practical implementation.

General speaking, the value of θ_c not only determines the cutoff frequency of the specified L_A , but also predicts the upper-stop bandwidth in the low-pass case and upper-pass bandwidth in the high-pass case. Fig. 2.5(a) plots the simulated frequency responses of the Chebyshev low-pass filter prototype with $\theta_c = 30^\circ$, 45° and 60° for $n = 2$ and $L_A =$

(a)

(b)

Fig. 2.5. Theoretical frequency responses under varied cutoff electrical length θ_c . (a) Low-pass filter prototype ($n = 2$). (b) High-pass filter prototype ($n = 2$).

0.1 dB. Obviously, the upper cutoff frequencies are entirely decided by $(180^\circ - \theta_c)$ and they are equal to 150° , 135° and 120° , respectively. In particular, frequency responses in this upper stopband are perfectly symmetric with respect to the 1st transmission zero at $\theta = 90^\circ$. In this aspect, the ratio $(180^\circ - \theta_c)/\theta_c$ is chosen as the frequency bandwidth of this upper stopband. Observing two inserted frames in Fig. 2.5(a) in the lower and upper

passband ranges of the stopband, the ripple levels are exactly equal to 0.1 dB for all three cases as specified in this synthesis design.

Similarly, frequency responses of the Chebyshev high-pass filter prototype can be calculated and plotted in Fig. 2.5(b) for $\theta_c = 30^\circ, 45^\circ$ and 60° under the same order $n = 2$ and the same ripple level $L_A = 0.1$ dB. In this case, the upper passband is also characterized as the relative bandwidth of $(180^\circ - \theta_c)/\theta_c$. As shown in Fig. 2.5(b), when θ_c increases, its upper cutoff counterpart $(180^\circ - \theta_c)$ decreases at the same speed. As such, an upper passband with symmetrical frequency response is realized and its bandwidth gradually becomes narrow as θ_c increases. Again, the ripple level in this high-pass band is exactly equal to 0.1 dB that is a specified value. Since its lower/upper cutoff frequencies can be explicitly determined, this constituted high-pass filter prototype can be applied to synthesis design of an alternative class of UWB filters.

Now, let us move to investigate the relative impedance ratios for different specified

Fig. 2.6. Relative stopband/passband bandwidth $(180^\circ - \theta_c)/\theta_c$ versus normalized impedance ratio b_1/a_1 for the low-/ high-pass filter prototypes with $n = 2$.

relative bandwidths. It is the reason for the range of impedance ratio is restricted in practical implementation using planar transmission-line structures. Fig. 2.6 shows the relative stopband/passband bandwidth $(180^\circ - \theta_c)/\theta_c$ versus impedance ratio b_1/a_1 for the low-pass and high-pass prototypes with $n = 2$ under different ripple levels, *i.e.*, $L_A = 0.1$ and 1.0 dB. For a low-pass prototype, its upper-stop bandwidth $(180^\circ - \theta_c)/\theta_c$ falls quickly from an extremely large value as b_1/a_1 increases in the range less than 1. Then, this bandwidth gradually becomes small to the value of 2. Therefore, the ratio b_1/a_1 must be chosen as small as possible if a low-pass filter with wide upper stopband needs to be designed. On the other hand, if a wide bandpass filter is constituted using the high-pass prototype, the ratio b_1/a_1 must be selected as a large value. Let us consider an UWB bandpass filter with a fractional bandwidth of 109.5%. Under 0.1 dB in-band ripple, its relative bandwidth is approximately equal to 3.42. From Fig. 2.6, it can be estimated that $b_1/a_1 = 1.77$. Generally speaking, the low-pass prototype filters ($n = 2, 3$) with $\theta_c = 30^\circ \sim 60^\circ$ under consideration are easier to be implemented in practice with reference to the port impedance of 50Ω and in-band ripple smaller than 1.0 dB. On the other hand, if two filters, either low-pass or high-pass, are designed under the same ripple level L_A and cutoff phase θ_c , but different order n , the one with higher order will have sharper rejection skirts or higher selectivity. In the next Section, two filters will be physically implemented on uniplanar transmission lines.

2.3.3 Filter implementations

Following the transmission-line models in Fig. 2.3(a) and (b), two filters with low-pass and high-pass performance are designed on hybrid slotline and coplanar-waveguide

Fig. 2.7. Compensation of cross-junction on slotline and coplanar-waveguide. (a) Geometry layout. (b) Frequency responses.

structures. In this aspect, one series stub with short-/open-circuited end is formulated by two slotline stubs in shunt. Hence, the normalized impedance of each slotline stub is twice as the original value listed in the tables in Appendix I and II. The uniform connecting lines between two adjacent stubs are realized by coplanar-waveguide lines. Different from the transmission-line models in Fig. 2.3(a) and (b), there exists a cross-

junction discontinuity at the connection point between slotline and coplanar-waveguide and it may affect the resultant performance of the realized filter circuits.

To minimize this unwanted parasitic effect, a compensated cross-junction structure is constructed with narrowed strip conductor width as shown in Fig. 2.7(a). Using the full-wave simulator HFSS [114], frequency responses of open-circuited series stubs with different sizes of compensation elements are numerically calculated and plotted in Fig. 2.7(b). If there is no compensation (strip width $s_{c1} = 1.0$ mm), the transmission pole occurs at 8.35 GHz. As s_{c1} is reduced to 0.70 mm and l_{c1} is increased, the pole gradually moves to 6.60 and 5.55 GHz. As such, this quantity can be used to tune the in-band high-pass performance. Similarly, it will also be employed to tune the frequency response with maximized insertion loss over the upper stopband for the low-pass filter.

In the implementation, two five-pole ($n = 3$) filters with low-pass and high-pass performance are designed and fabricated on a substrate with relative dielectric constant of 10.8 and thickness of 0.635 mm. The design parameters are selected from Appendix I with $\theta_c = 45^\circ$ for low-pass and $\theta_c = 40^\circ$ for high-pass. In particular, the high passband of the latter filter aims to achieve an ultra-wide passband from 3.1 to 10.6 GHz. The

TABLE 2.1
CIRCUIT PARAMETERS FOR LOW-PASS PROTOTYPE FILTER ($\theta_c = 45^\circ$ at 3 GHz)

Stub line impedance	Connecting line impedance
$Z_1 = Z_3 = 72.97 \, \Omega$	$Z_{1,2} = Z_{2,3} = 47.24 \, \Omega$
$Z_2 = 102.77 \, \Omega$	$Z_0 = 50 \, \Omega$

TABLE 2.2
CIRCUIT PARAMETERS FOR HIGH-PASS PROTOTYPE FILTER ($\theta_c = 40^\circ$ at 3 GHz)

Stub line impedance	Connecting line impedance
$Z_1 = Z_3 = 54.66 \, \Omega$	$Z_{1,2} = Z_{2,3} = 50.24 \, \Omega$
$Z_2 = 71.90 \, \Omega$	$Z_0 = 50 \, \Omega$

(a)

(b)

Fig. 2.8. Schematic layouts of two proposed prototype filters ($n = 3$). (a) Low-pass. Dimensions: $l_1 = l_3 = 5.82$, $l_2 = 6.47$, $l_{1,2} = l_{2,3} = 10.93$, $W_1 = W_3 = 0.34$, $W_2 = 1.0$, $s_{c1} = s_{c3} = 0.60$, $g_{c1} = g_{c3} = 0.58$, $l_{c1} = l_{c3} = 1.0$, $s_{c2} = 0.80$, $g_{c2} = 0.43$, $l_{c2} = 1.0$. (b) High-pass. Dimensions: $l_1 = l_3 = 5.10$, $l_2 = 5.20$, $l_{1,2} = l_{2,3} = 10.18$, $W_1 = W_3 = 0.11$, $W_2 = 0.33$, $s_{c1} = s_{c3} = 0.74$, $g_{c1} = g_{c3} = 0.51$, $l_{c1} = l_{c3} = 1.0$, $s_{c2} = 0.70$, $g_{c2} = 0.53$, $l_{c2} = 1.0$. All units are in mm

impedance values of the designed low-pass and high-pass filters are tabulated in Table 2.1 and 2.2, where Z_i and $Z_{i,i+1}$ are the characteristic impedances of the i^{th} series stub and that of the line connecting the i^{th} and $(i+1)^{\text{th}}$ series stubs. To this end, all the initial dimensions of these short-/open-circuited stubs and connecting lines can be derived via the proposed direct synthesis procedure in the above Section. Due to the unexpected dynamic and discontinuity effects at high frequencies as also pointed out before, the final overall dimensions of these two filter layouts including compensated cross-junction [16],

(a)

(b)

(c)

Fig. 2.9. Low-pass filter ($n = 3$) using hybrid slotline and coplanar-waveguide structures. (a) Photograph of the fabricated filter. (b) Predicted and measured S-parameters. (c) Predicted and measured group delays.

are optimally designed via full-wave ADS [115] and HFSS simulators. Fig. 2.8(a) and (b) depict the final schematic layouts of these two filters with all the dimensions labelled.

Next, these two filter circuits are fabricated and their relevant photographs are illustrated in Fig. 2.9(a) and 2.10(a), respectively. The measured frequency responses are shown in Fig. 2.9(b) and 2.10(b) against the predicted ones based on transmission-line theory and full-wave simulation. Three sets of in-band group delays obtained from three different approaches for the low-pass and high-pass are also depicted in Fig. 2.9(c) and 2.10(c). All the three sets of frequency responses and group delays are found in good agreement with each other over the plotted frequency range. More detailed comparisons

(a)

(b)

Fig. 2.10. High-pass filter ($n = 3$) using hybrid slotline and coplanar-waveguide structures. (a) Photograph of the fabricated filter. (b) Predicted and measured S-parameters. (c) Predicted and measured group delays.

of electrical parameters among these three sets of results are given in Table 2.3 and 2.4 for the low-pass and high-pass filters, respectively. In these tables, 3 dB cutoff frequencies are used to do the comparative studies instead of 0.1 dB ripple level point since 3-dB bandwidth has been widely used in design of almost all of filter for wireless communication. For the low-pass filter, three consistent transmission poles emerge in the low passband less than about 3.4 GHz while a transmission zero with deep attenuation appears around 5.8 to 6.0 GHz. For the designed high-pass or wide band-pass filter, the five transmission poles are observed in the realized wide passband where the return losses for all the three cases are higher than 10 dB. The in-band group delays for both the low-pass and high-pass filters are very flat and become larger near the cutoff frequencies as a trade-off of the high selectivity.

Although, the predicted and measured results agree with each other well, it can still be seen that some visible discrepancies among the three sets of results, which are obtained from three different approaches. Particularly at high frequencies, some visible shifts in frequency or bandwidth between the predicted and measured results can be

found and they are primarily caused by many unexpected factors such as inaccurate and frequency-dispersive permittivity (10.8 ± 0.25 at 10 GHz given in the datasheet), as well as finite-thickness of the strip conductor and the tolerance in etching process.

TABLE 2.3
COMPARISONS FOR LOW-PASS PROTOTYPE FILTER

	Theory	EM simulation	Measurement
3 dB cut-off (GHz)	3.44	3.45	3.34
Maximum in-band insertion loss (dB)	0.10	0.12	0.54
Minimum in-band return loss (dB)	16.43	17.72	15.51
3 dB upper cut-off (GHz)	8.56	8.12	7.89
Maximum in-band group delay variation (ns)	0.19	0.18	0.24

TABLE 2.4
COMPARISONS FOR HIGH-PASS PROTOTYPE FILTER

	Theory	EM simulation	Measurement
3 dB cut-off (GHz)	2.43	2.37	2.41
Maximum in-band insertion loss (dB)	0.10	1.55	2.28
Minimum in-band return loss (dB)	16.43	18.73	10.61
3 dB upper cut-off (GHz)	11.06	11.21	10.40
Maximum in-band group delay variation (ns)	0.14	0.09	0.23

To take the effect of the fabrication tolerance into account in the filter’s performances, three sets of filters have been simultaneously fabricated in one bench for our designed filters through this dissertation. As told by our vender, the minimum strip/slot width that can be fabricated is about 0.1 mm. This tolerance may be combined with the other critical dimensions such as conductor thickness and width so as to affect the filter performance. To quantitatively demonstrate this problem, three curves with the

Fig. 2.11. Measured S-parameters of low-pass filter shown in Fig. 2.9(a) with added tolerances. Filter I: $l_1 = l_3 = 5.72$, $l_2 = 6.37$, Filter II: $l_1 = l_3 = 5.82$, $l_2 = 6.47$, Filter III: $l_1 = l_3 = 5.92$, $l_2 = 6.57$. All units are in mm

consideration of ± 0.1 mm in tolerance are experimentally derived for the low-pass filter in Fig. 2.9(a) as shown in Fig. 2.11. From this figure, it can be observed that the positions of transmission zeros are slightly different due to the variation in tolerance. However, the passband responses are almost the same as each other. By comparing these measured results with the simulated results, it is noticed that the position of the transmission zero in the simulation is still at higher frequency. Furthermore, it is demonstrated that the measured results of the 3rd filter (Filter III) that is designed within the range of tolerance is now closer to the simulated one.

On the other hand, as mentioned above, the inaccurate or frequency-dependent dielectric constant can also cause the difference between the measured and simulated results. In the full-wave simulation, the dielectric constant used is 10.8 and it is a constant throughout the simulation frequency range. However, the actual material may have a

Fig. 2.12. Simulated S-parameters of low-pass filter shown in Fig. 2.9(a) with different dielectric constants.

± 0.25 or even larger inaccuracy which can influence the frequency responses. To show the effect of dielectric constant variation, three sets of low-pass filters with the same dimensions and varied dielectric constants ϵ_r are simulated and their S-parameters are plotted in Fig. 2.12. From the figure, it can be observed that the return loss level as well as the position of transmission zero are both affected by the change of dielectric constant. There is obviously frequency shift among these three sets of curves. Therefore, the variation of dielectric constant of the substrate may be one of the other reasons that make the measured and simulated results look differently.

In addition, for the high-pass case shown in Fig. 2.10(b), the transmission-line based theoretical S-parameter magnitudes, *i.e.*, $|S_{11}|$ and $|S_{21}|$, behave much more differently than the simulated and measured ones over the frequency range DC to 2.4 GHz. In fact, the open-circuited slotline stubs should be considered as a coupling gap at low frequencies and this gap capacitive effect cannot be accurately included in the equivalent transmission-line model as shown in Fig. 2.3(b). As a consequence, the measured and

simulated $|S_{21}|$ fall much more slowly than that based on the transmission-line theory as the frequency is decreased to DC.

To further understand loss behavior of the two designed filter structures on uniplanar slotline and coplanar waveguide, two relevant sets of frequency-dependent loss factor ($1 - |S_{11}|^2 - |S_{21}|^2$) are calculated and illustrated in Fig. 2.13(a) and (b). From the full-wave simulation and measurement under two different conditions, *i.e.*, perfect or copper strip conductor and lossless or lossy dielectric substrate, it can be figured out that the radiation loss occupies a large portion of the overall loss. For the low-pass filter, the loss gradually rises up as the frequency increases and reaches to its maximum around 8.0 GHz, where the transverse slots become resonant. For the high-pass filter, its first and second peaks in loss appear around 2.5 GHz and 10.0 GHz, *i.e.*, the lower and upper cut-off frequencies of the constituted high passband. This loss may be suppressed by putting these filters in a metallic enclosure if necessary at the cost of enlarged size, heavy weight and expensive fabrication. In addition, it will also bring in the cavity modes of the box that depend on the dimension of the enclosure, and they will interfere with the filter performance.

The implemented low-/high-pass filters suffer from the higher-order harmonic bands due to the frequency-distributed property of the transmission-line structures. In order to suppress the harmonic bands and extend the upper stopband, periodic structures are used to replace the uniform connecting lines between the adjacent stubs. The periodic structures have similar electric properties, *i.e.*, characteristic impedance and propagation constant, as the original uniform lines within the concerned passband. However, the attenuation factor becomes nonzero in the harmonic bands. Hence, the harmonic bands will be suppressed and the upper stopband will be extended [116]-[117].

(a)

(b)

Fig. 2.13. Simulated and measured loss factors versus frequency. (a) Low-pass filter in Fig. 2.9. (b) High-pass filter in Fig. 2.10.

A low-pass filter with widened upper stopband is designed and fabricated. Since the propagation constant for the periodic lines increases near the designed cutoff frequency, it affects the performance of the first harmonic band of the low-pass filter. Therefore, additional periodic structures are used at two ports with characteristic impedance as 50Ω to enhance the suppression at the first harmonic band edge. The photograph of the fabricated low-pass filter with upper stopband extension is shown in Fig. 2.14(a). The implemented filter is measured and its frequency responses are plotted together with the previous low-pass filter without harmonic suppression in Fig. 2.14(b). It can be observed that the performance in the desired low passband has been almost unchanged, while the

(a)

(b)

Fig. 2.14. (a) Photograph of low-pass filter with harmonic bands suppression. (b) Measured S-parameters comparison between low-pass filter without harmonic suppression (dotted line) and low-pass filter with harmonic suppression (solid line).

upper stopband has been greatly extended to 15 GHz with a suppression level below 16.7 dB.

There is an alternative implementation structure of the proposed low-/high-pass filters using series stubs shown in Fig. 2.3. Again, the series stubs are realized by slotlines with short-/open-circuited ends on the ground. The connecting lines that cascade the adjacent stubs are implemented with microstrip lines on the top of the substrate. The

transition between microstrip line and slotline is modeled as an ideal transformer with ratio 1 due to the thin substrate used here [16]. A low-pass filter is designed as an example to illustrate this kind of implementation using hybrid microstrip and slotline. The design specifications are the same as in the previous example that are listed in Table 2.1 and the schematic layout is shown in Fig. 2.15. At the cross-junction of microstrip line and slotline, the narrowed microstrip line width is also used to compensate the junction effect. From Fig. 2.15, it is noticed that the central cross-junction is not compensated. As mentioned above, the narrowed-strip is mainly used to compensate the series capacitance brought in by the slotline stubs, especially at low frequency. Since the central slotline is wider, the capacitance is smaller. Hence, the central cross-junction does not need as much compensation as two side junctions.

The low-pass filter depicted in Fig. 2.15 is fabricated and measured, the measured frequency responses are plotted together with the ones obtained from full-wave

Fig. 2.15. Schematic layouts of proposed low-pass prototype filter ($n = 3$) implemented with hybrid microstrip line and slotline structures. Dimensions: $l_1 = l_3 = 5.91$, $l_2 = 6.50$, $l_{1,2} = l_{2,3} = 8.89$, $W_{1,2} = W_{2,3} = 0.33$, $W_1 = W_3 = 0.61$, $W_2 = 1.0$, $l_{c1} = l_{c3} = 1.0$, $w_{c1} = w_{c3} = 0.38$. All units are in mm

Fig. 2.16. Low-pass filter ($n = 3$) using hybrid microstrip line and slotline structures. (a) Photograph of the fabricated filter. (b) Predicted and measured S-parameters.

simulation and transmission line theory in Fig. 2.16. The three sets of curves agree with each other very well, especially in the lower passband. The position of the transmission zero in the measured results shifts to the left from 6 GHz to 5.50 GHz, which is probably due to the short-end effect of the series slotline stubs. This structure provides alternative implementation scheme for the proposed low-pass filter prototype. Similarly, the transmission-line model of the high-pass filter depicted in Fig. 2.3(b) can also be realized

using hybrid microstrip line and slotline structures. However, the open ends of the series stubs have to be implemented by cutting the substrate out.

2.4 Conclusion

In this chapter, a class of transmission line low-/high-pass filter prototypes with short-/open-circuited series stubs has been presented and studied. A direct synthesis approach has been established for the optimization design of these prototypes with Chebyshev equal-ripple characteristics. After a transfer function is introduced, the element values in the prototypes, *i.e.*, normalized characteristic impedances of the series stubs and connecting lines, are analytically determined and tabulated for different orders and ripple levels. Two filter blocks are then implemented both on the hybrid slotline and coplanar-waveguide structures and hybrid microstrip line and slotline structures. To improve the upper stopband performance of the proposed low-pass filter prototype, periodic structures are used to replace the uniform lines in the original transmission-line model. With the slow-wave behavior of the periodic lines, the upper stopband can be efficiently extended without affecting the low passband performance. The predicted results on these simple transmission-line models are validated by full-wave simulation and measurement. The proposed filter prototypes provide us with an effective methodology in constructing a low-pass filter with wide upper stopband and a band-pass filter with the specified ultra-wide bandwidth. The advantage of the proposed filter prototype with series stubs over the conventional high-pass prototype with shunt stubs is that it can provide an alternative way to design low-/high-pass filters on hybrid planar transmission line structures and can realize some desired bandwidths that cannot be easily

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achieved by the traditional high-pass prototype. Especially, for the high-pass or quasi-bandpass filter, the proposed approach can avoid the use of via holes in practical implementation, thus it makes the fabrication process simpler.

CHAPTER 3

UWB Bandpass Filters using Stub-/Slot-Loaded MMRs

3.1 Introduction

Beside the conventional high-pass prototype with shunt short-circuited stubs and its dual network with series open-circuited stubs introduced in Chapter 2, there are several other ways to design UWB bandpass filters. As mentioned earlier, a primary class of UWB bandpass filters that are designed by using MMR becomes quite popular recently due to its simple design procedure, compact size and good performance [57]-[72]. The concept of building UWB bandpass filter with MMR was initially presented in [57]. Unlike the traditional parallel-coupled resonator filters using single resonance [118], this type of filters utilizes first few resonant modes of a MMR to construct the filter passband. By properly adjusting the impedance ratio of the MMR, the resonant frequencies are able

to be located within the targeted passband. With the coupling structures at two ports, a passband with ultra-wide bandwidth can be established. At the same time, the bandwidth of the filters can be controlled by the designers. The geometrical layout and the corresponding transmission-line model of a basic open-ended MMR implemented on microstrip line structure are depicted in Fig. 3.1(a) and (b), respectively.

As shown in Fig. 3.1(a), the MMR consists of two identical transmission-line sections with characteristic impedance Z_1 and electrical length θ_1 at two sides and one

Fig. 3.1. Open-ended microstrip line MMR. (a) Schematic layout. (b) Transmission-line model.

section with impedance Z_2 and electrical length $2\theta_2$ in the middle. Here it is assumed that Z_1 is larger than Z_2 . From the equivalent transmission-line mode of this MMR, the input admittance Y_{in} looking from one open-end can be derived as,

$$Y_{in} = jY_1 \frac{2(R \tan \theta_2 + \tan \theta_1)(R - \tan \theta_1 \tan \theta_2)}{R(1 - \tan^2 \theta_1)((1 - \tan^2 \theta_2) - 2(1 + R^2) \tan \theta_1 \tan \theta_2)} \quad (3.1)$$

where R is the impedance ratio and $R = Z_1/Z_2$. The resonance condition of such a resonator is that,

$$Y_{in} = 0 \quad (3.2)$$

To make things simpler, here we assume $\theta_1 = \theta_2 = \theta$. The resonant frequencies f_1 , f_2 and f_3 can be determined as,

$$\theta(f_1) = \tan^{-1} \sqrt{R} \quad (3.3a)$$

$$\theta(f_2) = \frac{\pi}{2} \quad (3.3b)$$

$$\theta(f_3) = \pi - \tan^{-1} \sqrt{R} \quad (3.3c)$$

Therefore, it can be observed from above equations that the relative locations of spurious resonances with respect to the fundamental frequency are determined by the impedance ratio. By properly adjusting the impedance ratio, the first two, three and even four resonant frequencies can be squeezed together to construct the passband of an UWB filter with the help of feeding structures. In [57], the MMR is simply feed by parallel-coupled lines at input/output ports.

Recently, *Zhewang Ma* and his colleagues have done a lot of work in developing a synthesis method to design MMR UWB filters [119]-[123]. The parallel-coupled transmission lines are represented with an equivalent circuit with two open-circuited stubs connected through a uniform transmission line [19]. The schematic layout and equivalent circuit model of the parallel-coupled microstrip line are depicted in Fig. 3.2 (a) and (b). By replacing the parallel-coupled lines in [57] with the equivalent circuit in Fig.

Fig. 3.2. Parallel-coupled microstrip line. (a) Schematic layout. (b) Equivalent transmission-line model.

Fig. 3.3. Transmission-line model of the MMR UWB filter in [57].

3.2(a), the transmission-line model of the basic MMR UWB bandpass filter is established in Fig. 3.3.

By observing this transmission-line network, it is noticed that it is similar with the one in Fig. 2.3(b) except the electrical length of connecting line between the first and second open-circuited stubs is θ_c instead of $2\theta_c$. The analysis method is the same as that introduced in Chapter 2, so that the unknown element values in the network can be calculated. The dimensions of the designed filter are able to be determined consequently.

3.2 UWB Bandpass Filter using Stub-Loaded MMR

Another way to design this kind of UWB bandpass filters with MMR is to use mode theory. For an open-circuited half wavelength MMR, the current distribution along the resonator can be represented by sine waves as depicted in Fig. 3.4. The minimum values indicate an open circuit, while the maximum values indicate a short circuit. Hence, the central plane of a line resonator represents a perfect electric wall (P.E.W.) or short circuit for all the odd-order modes of the line resonator and a perfect magnetic wall (P.M.W.) or

Fig. 3.4. Current distributions of an open-circuited uniform transmission-line resonator. $n = 1, 2, 3, 4$ ($l = \lambda/2, \lambda, 3\lambda/2, 2\lambda$).

open circuit for all the even-order modes. Therefore, loading an open-circuited stub in shunt at the center position will not affect the odd-order resonances but influences the even-order resonances greatly.

As introduced in last section, no matter the initial MMR UWB bandpass filter utilizing the first three resonance modes to realize a desired ultra-wide passband [57] or the later synthesis designed UWB bandpass filter with single-stage and multiple-stage resonators [122], this kind of filter usually suffers a narrow upper stopband due to the next unused resonant modes of the resonator. Take the UWB bandpass filter in [57] as an example. By viewing the resonant frequency responses graphically under weak capacitive coupling gaps at two ports [62], it can be observed that the fourth resonant frequency occurs at around 13.2 GHz, which leads to a relatively narrow upper stopband of 11.0 to 14.0 GHz.

In order to achieve a wider upper stopband, the fourth resonant frequency becomes the main concern. Although it can be eliminated by some means, such as capacitive-ended interdigital coupling at two ports [62], it can also be utilized to construct the passband together with the first three resonances. As mentioned earlier, loading an open-ended stub in shunt at the central plane of the line resonator will affect the even-order resonant modes but not the odd-order modes. Similarly, the planes at an approximately one third wavelength at central frequency from both ends indicate perfect magnetic wall or open circuit for the third resonant frequency. Therefore, loading open-ended stubs in shunt at that position will affect the third resonant mode. But, at the same time, have little effects on the other modes. So there are two additional degrees of freedom to adjust the first four resonant frequencies to allocate them evenly within the UWB passband. With

this idea in mind, the stub-loaded MMR is proposed and its schematic layout is depicted in Fig. 3.5.

Fig. 3.5. Schematic of the novel stub-loaded MMR.

Loading open-ended stubs in shunt at those specific places will lengthen the effective surface current length of the corresponding resonant mode; these modes will move forward gradually with the increasing stub lengths. The following figures illustrate the effect of the loading stubs. As shown in Fig. 3.6, the length of the central stub is indicated by L_C and the length of the stubs at two sides is indicated by L_S . Firstly, to study the effect of the central stub to the even-order resonant frequencies, L_S is fixed to a constant value and L_C is increased gradually from 1.8 mm to 2.3 mm and 2.8 mm. It can be observed from Figure 3.6(a), the second and the fourth resonant frequencies reduce slowly and quickly with the increasing L_C , while the first and the third resonances are almost unchanged. This phenomenon verifies what has been discussed earlier. On the other hand, fix $L_C = 2.29\text{mm}$ and vary L_S from 1.7mm to 2.2mm and 2.7mm, the third resonant frequency drops from 9.47 to 8.76 and 8.03 GHz as depicted in Fig. 3.6(b). As a result, the second, third and fourth resonances can be easily controlled by the lengths of the loading stubs. By proper choosing the dimensions, the first four resonant frequencies are all involved in contribution to the UWB passband.

(a)

(b)

Fig. 3.6. Frequency-dependent $|S_{21}|$ of the stub-loaded MMR versus lengths of the loaded stubs. (a) Central-stub length L_c . (b) Side-stub length L_s .

To realize the filtering performance of the resonator, strong coupling is needed at input and output ports. Parallel-coupled lines provide a relatively weak coupling, as used in [57], the gap width of 0.05mm between the parallel-coupled lines, can only achieve -10dB in-band return loss. Therefore, interdigital coupled lines with increased coupling strength are installed at the two ports of the proposed UWB stub-loaded MMR to form the bandpass filter. In this case, the fabrication difficulty is lowered since the gaps between the parallel lines are all widened to 0.1mm, which is twice compared with that in [57]. At the same time, the in-band return loss is higher than 15 dB in the simulated

Fig. 3.7. Schematic layout of the proposed stub-loaded MMR UWB bandpass filter. Substrate: Roger's RT/Duriod 6010, $\epsilon_r = 10.8$ and $h = 1.27$ mm.

results. After some optimization works, the final proposed UWB bandpass filter is designed and EM full-wave simulation is done. The configuration of the optimized UWB bandpass filter with the dimensions labeled is shown in Fig. 3.7.

The simulated results show a good UWB bandpass filtering performance with -3dB bandwidth from 2.98 GHz to 10.49 GHz, insertion loss lower than 1.3 dB, return loss higher than 15 dB and upper stopband up to 15 GHz. To verify the simulation results, a prototype UWB bandpass filter was fabricated and measured. The material used is Roger's RT/Duriod 6010 with relative permittivity: $\epsilon_r = 10.8$ and thickness: $h = 1.27$ mm. The photo of the fabricated filter is shown in Fig. 3.8 (a). In order to eliminate the effects

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of improper soldering and provide more accurate results, universal test fixture was used in the measurement. By employing the Agilent Network Analyzer and the universal test fixture, the measurement of S-parameters of the fabricated UWB bandpass filter sample can be done.

The comparisons between the simulated and measured S_{11} - and S_{21} -magnitudes, as well as group delay in a frequency range from 1 GHz to 15 GHz are shown in Fig. 3.8(b)

(a)

(b)

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Fig. 3.8. Stub-loaded MMR UWB bandpass filter. (a) Photograph. (b) Simulated and measured S-parameters. (c) Simulated and measured group delays.

and (c), respectively. First of all, it can be seen that these simulated and measured results are in very reasonable agreement with each other. Some minor discrepancies between them may be caused by unexpected tolerances in fabrication and material parameters. As investigated in Chapter 2, the inaccurate fabrication or substrate parameter variation can cause frequency shift as observed from Fig. 3.8(b). The lower and higher cutoff frequencies of the UWB passband in the measurement are equal to 2.80 and 10.27 GHz against their counterpart frequencies of 2.98 and 10.49 in the simulation. Furthermore, the fabricated UWB filter has achieved the return loss higher than 14.3 dB over the realized UWB passband with a fractional bandwidth of 114%.

As compared to the initial filter in [57], this filter gets the increased roll-off in the upper cutoff frequency. In the meantime, the simulated and measured group delays within the UWB passband are found less than 0.87 and 0.81 ns, respectively, with the maximum variation of 0.64 ns. This raised variation is caused by a trade-off between rejection skirt in S_{21} -magnitude and linearity in phase as concluded in [113]. In final, it can be found that the overall length of this proposed UWB bandpass filter is 13.80 mm, smaller than

15.64 mm of the initial UWB filter in [57]. It is benefited by the enhanced slow-wave factor of stub-loaded structure in the central part of the proposed MMR.

In [124], Chiou and Kuo investigated where the fourth resonance of our proposed stub-loaded MMR in Fig. 3.5 and the transmission zero at the upper passband edge shown in Fig. 3.8(b) come from. After studying the resonant frequency responses for two cases, i.e., $L_C = 0$ and $L_S = 0$, they concluded that “the fourth resonance is one of the inherent resonant modes of the unloaded MMR” and the transmission zero is introduced by the loaded stubs. With full agreement of their comments, there are some discussions below that further illustrate this proposed stub-loaded MMR UWB bandpass filter.

As mentioned in the comment [124], when $L_C = 0$ is selected and L_S is increased, all resonances move down to lower frequencies as shown in Fig. 3.6(b). Among them, the third resonance f_3 and the fourth resonance f_4 are reduced more rapidly than first two resonant frequencies, i.e., f_1 and f_2 . One reason is that the loading positions of the side stubs (L_S) are designated at the open-circuited planes of the third resonant frequency, which are close to the open-circuited planes of the fourth resonance. Therefore, the variation of L_S influences these two resonances much more than the other two. The other reason is that the increase in effective current length as a function of frequency more significantly affects the variation on higher resonances since these stubs have shorter wavelengths at low frequencies. In Fig. 1(b) of the comment [124], it is interesting to notice that f_1, f_2 and f_z intersect at one point when $L_C / (\lambda / 4) \approx 1$ as also found in [125]. As their comment says, the frequency point of $f_1 = f_z$ shows that the transmission zero is introduced by the loading stub, but why f_2 does join them? It can be seen that when the central stub length $L_C \approx \lambda / 4$, the effective current length of the second resonance

becomes double. As a result, f_2 is reduced by half and equals f_1 . This result can be also used to explain why f_4 is nearly twice of f_1 at $L_C / (\lambda / 4) \approx 1$.

3.3 Wideband Bandpass Filter using Slot-Loaded MMR

3.3.1 Single-stage wide-band bandpass filter with 59% fractional bandwidth

Similarly, bandpass filters with wide or even ultra-wide bandwidth can be designed using slot-loaded MMR. Starting from the MMR shown in Fig. 3.1, a mode-graph can be plotted versus the width of the central low-impedance transmission-line section (W_2) with different length ratio u of the two sections with different characteristic impedances. For fabrication accuracy consideration, the width of the high-impedance transmission-line sections at two sides is fixed as 0.2 mm

$$u = \frac{L_1}{L_1 + L_2} \quad (3.1)$$

where L_1 and L_2 are physical lengths of two distinctive microstrip sections with different characteristic impedances Z_1 and Z_2 . It can be observed from the figure that, when the line resonator is uniform ($W_2 = W_1 = 0.2$ mm), the resonant frequencies all occur at integer multiple times of the fundamental frequency f_1 . Meanwhile, the resonant frequencies of a non-uniform resonator depend on the impedance ratio as well as the length ratio. For a fixed length ratio 0.5, the first three resonant frequencies squeeze together as W_2 increases, at the same time the fourth resonant frequency f_4 drops down. When u is smaller than 0.5, f_2 and f_3 are moving closer with the increased W_2 , and the spacing between the first and the second resonant frequencies becomes larger. It is

interesting to see that in the $u = 0.53$ case, the fourth resonant frequency keeps almost unchanged with the increasing W_2 , which has the potential to design the wide-band bandpass filter with relatively wide upper stopband if we could use the first three modes.

To design the bandpass filter with a wide bandwidth of about 50% to 70% or even larger, the first three modes should be better all involved to form the passband. In the case of $u = 0.53$ as shown in Fig. 3.9, the first two resonances move closer, while the third resonance is far away from them when W_2 is increased. If the third resonance can be moved forward while keeping the first two modes unchanged at the same time, a passband can be constructed by the first three resonant frequencies with proper external coupling. Similar mode theory used in last section can be applied in this situation. Since the central plane of the MMR resonator indicates the open circuit for the even modes of the MMR, etching two identical slits from the top and bottom of the resonator to the

Fig. 3.9. Resonant spectrum of SIRs with $u = 0.47$ (dotted lines), 0.5 (dashed lines) and 0.53 (solid lines). Substrate: Roger's RT/Duriod 6010, $\epsilon_r = 10.8$ and thickness: $h = 1.27$ mm.

transverse center at the longitudinally central plane will not affect the status of all the even-order resonant modes. However, it will increase the effective current length of the odd modes, and move them forward. Based on this idea, a novel slot-loaded stepped-

Fig. 3.10. Schematic layout of proposed slot-loaded MMR. Dimension: $L_1 = 6.29$ mm, $L_2 = 2.79$ mm, $W_1 = 0.2$ mm, $W_2 = 8$ mm.

Fig. 3.11. Frequency-dependent $|S_{21}|$ of the slot-loaded MMR versus central-slot length d .

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impedance MMR is presented and its geometry is shown in Fig. 3.10. The lengths and widths of the two sections with different characteristic impedances are indicated by L_1 and L_2 , W_1 and W_2 , respectively. The length of the loaded slots is labeled as d and the width of the slots is fixed as 0.2 mm for easy fabrication.

To illustrate the effect of the slots, the resonant frequency responses against different slot lengths d are plotted in Fig. 3.11. From the figure, it can be seen that as d varies from 0 mm, to 1.9 mm and 3.8 mm, the first and the third resonances are moving forward slowly and quickly, however, the second and the fourth resonant frequencies are almost unchanged as predicted. It needs to be pointed out that the resonance occurrence near 10.2 GHz is excited from transverse direction due to the enlarged W_2 . Once the third resonant frequency moves close enough, the second resonance can be located at the middle of the first and the third resonances, thus constructing a wide-band bandpass filter using the first three modes. Parallel coupled lines are used at the input and the output ports, and the schematic of an optimized bandpass filter with dimensions indicated is shown in Fig. 3.12.

Fig. 3.12. Schematic of slot-loaded wide-band bandpass filter.

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The designed filter is fabricated and measured. The photograph of this fabricated filter is shown in Fig. 3.13(a), and the measured frequency responses and group delay are plotted together with those obtained from simulation in Fig. 3.13(b) and (c), respectively. This filter has achieved a wide bandwidth from 3.0 GHz to 5.6 GHz (59% fractional bandwidth) with insertion loss less than 1.0 dB, return loss higher than 15.0 dB and upper stopband up to 11.8 GHz ($|S_{21}|$ below -10dB). The group delay is only plotted from 1 GHz to 8 GHz, because the group delay within the passband is the main concern. From the figure, it can be observed that the maximum variation within the passband is 0.23 ns,

(a)

(b)

Fig. 3.13. Slot-loaded wide-band bandpass filter. (a) Photograph. (b) Simulated and measured S-parameters. (c) Simulated and measured group delays.

which indicates good phase linearity. The concept of lower and upper bands of UWB becomes popular to avoid the interferences of existing radio systems to UWB communications, such as WLAN. This filter has the potential application to be implemented in the design of lower UWB band bandpass filter.

3.3.2 Two-stage wideband bandpass filter with 60% fractional bandwidth

Selectivity of this wide-band bandpass filter can be improved by cascading more sections of individual MMR [90]–[91]; however, the trade-off is that it increases the total circuitry size and insertion loss. A six-pole bandpass filter can be constructed by using two identical slot-loaded MMRs. The split modes graph is plotted in Fig. 3.14 to investigate the split off of the resonant peaks of cascading MMRs with respect to coupling gap size S under weak coupling at the input/output stage. The cascaded two sections can also be treated as one resonator with six resonant frequencies varying with the coupling length between them. The coupling length between two MMRs is fixed at

Fig. 3.14. Mode graph of the two cascaded slot-loaded MMRs.

around one quarter guided wavelength at center frequency. The split-off increases when S is decreased and $S = 0.6$ mm is chosen for the two-stage case to place the six resonances with even in-between distance.

The six-pole wide-band bandpass filter is realized by applying parallel coupling at the two ports. The two-stage wideband bandpass filter is then fabricated and measured. The photograph of the fabricated circuit is shown in Fig. 3.15 (a), and the measured S-parameters and group delay are plotted together with those simulated ones in Fig. 3.15(b) and (c), respectively. It can be observed from the figures that the simulated and measured frequency responses and group delays are in good agreement. It shows a 3-dB passband from 3.02 GHz to 5.63 GHz, *i.e.*, 60.3% fractional bandwidth at 4.33 GHz, mid-band insertion loss less than 1.24 dB and return loss is greater than 10 dB within the main passband. The in-band maximum group delay variation is 1.38 ns. As compared with

(a)

(b)

(c)

Fig. 3.15. Two-stage slot-loaded wide-band bandpass filter. (a) Photograph. (b) Simulated and measured S-parameters. (c) Simulated and measured group delay.

single-stage case, the filter selectivity is effectively enhanced and the upper stopband performance is improved.

3.3.3 Single-stage T-slot UWB bandpass filter with 110% fractional bandwidth

This slot-loaded MMR structure is not only capable of designing wide-band bandpass filter, it is also adopted to design filter with ultra-wide bandwidth with fractional bandwidth over 100%. Fig. 3.16 (a) shows the evolution of the proposed T-slot MMR. To miniaturize the slot-loaded MMR, W_1 is chosen as the smallest strip width that

Fig. 3.16. Evolution of the proposed T-slot MMR. (a) Schematics. (b) Resonant frequency responses. Dimension: $L_1 = 4.73$, $W_1 = 0.1$, $L_2 = 2.03$, $W_2 = 1.34$. All units are in mm. Substrate: Roger’s RT/Duriod 6010, $\epsilon_r = 10.8$ and $h = 1.27$ mm.

can be fabricated with ensured accuracy. L_1 , L_2 as well as W_2 are selected such that the central frequency is located around 6.85 GHz which is the center of the unlicensed UWB band. Similar with the slot-loaded MMR in previous section, two slots are formed in the central plane of the basic MMR I shown on the top. Therefore, as the length of vertical slot t_1 increases, the first and third resonances will move forward. However, when t_1 becomes larger and close to half of the width of the middle low-impedance line section as the second MMR in Fig. 3.16(a), the third resonant mode is still not close enough to the second resonant mode. Hence, the length of the slot has to be increased further more and it expands horizontally shown as MMR III.

To verify and validate the above theory, the resonant frequencies of these three MMRs under weak coupling at two ports are plotted in Fig. 3.16(b). As expected, the first and third resonances are moving from 4.76 GHz, 4.65 GHz to 4.35 GHz and 11.30 GHz, 10.81 GHz to 9.48 GHz as the total length of the slot ($t_1 + 2t_2$) increases. At the same time, the second and fourth resonant frequencies are almost unchanged. When the first and third resonances move to the positions that the distances to the designed central frequency are approximately the same, the UWB passband can be constructed by using the first three modes under the strong interdigital coupling at input and output ports. The schematic layout of the constructed filter is shown in Fig. 3.17(a) with all dimensions labeled. The implemented filter is then fabricated and measured. Fig. 3.17(c) plots the simulated and measured S-parameters of the T-slot MMR UWB filter, and it shows good agreement between these two sets of curves. An UWB bandpass filter with 110.27% FBW (3.10 GHz – 10.72 GHz) at 6.91 GHz is realized, and its maximum group delay variation within the passband is 0.14 ns. The noticeable shift in the location of the fourth

Fig. 3.17. Single-stage T-slot MMR UWB bandpass filter. (a) Schematic. (b) Photograph. (c) Simulated and measured S-parameters. (d) Simulated and measured group delay.

resonant frequency between the simulation and measurement may be caused by the

tolerance in fabricating the interdigital coupling lines and discontinuity effect in measurement with virtue to SMA connector or test fixture. But, the return loss in both simulation and measurement has exceeded 20dB around this resonance, showing their good agreement with each other.

3.3.4 Two-stage T-slot UWB bandpass filter with 110% fractional bandwidth

With the help of mode-graph, two-stage T-slot MMR UWB bandpass filter is

(a)

(b)

(c)

Fig. 3.18. Two-stage T-slot MMR UWB bandpass filter. (a) Photograph. (b) Simulated and measured S-parameters. (c) Simulated and measured group delay. Substrate: Roger’s RT/Duriod 6010, $\epsilon_r = 10.8$ and $h = 1.27$ mm.

designed and fabricated on the same substrate. The dimension of the two T-slot loaded MMRs is the same as that in Fig. 3.17(a). The rejection skirts at lower-/higher-passband edges are improved significantly and the upper stopband is also extended. The photograph of the fabricated filter is shown in Fig. 3.18(a), and the simulated and measured S-parameters and group delays are depicted in Fig. 3.18(b) and (c), respectively. The measured frequency responses exhibit a good UWB filtering performance from 3.15 GHz to 10.81 GHz, *i.e.*, 109.7% FBW at 6.98 GHz. The mid-band insertion loss is 1.45 dB and the return loss within the whole passband is below 11 dB. The upper stopband is up to 18.5 GHz with 20 dB suppression and the maximum group delay variation within the passband is 0.33 ns.

3.3.5 Single-stage T-slot UWB bandpass filter with enhanced selectivity

Besides increasing the stage number, filter selectivity can also be improved by purposely introducing two transmission zeros at lower and upper passband edges,

Fig. 3.19. Transmission-line model of a two-section open-circuited stub.

respectively. The transmission zeros can be generated by two open-circuited stubs loading at the two ports. The transmission-line model of this stub is depicted in Fig. 3.19. As shown in the figure, it is composed of two sections with electrical length θ_1 and θ_2 , characteristic impedances Z_1 and Z_2 that are used to control the locations of the two transmission zeros. The input admittance looking from one end of the open-ended stub can be derived as following,

$$Y_{in} = jY_1 \frac{Y_2 \tan \theta_2 + Y_1 \tan \theta_1}{Y_1 - Y_2 \tan \theta_1 \tan \theta_2} \tag{3.2}$$

If such an open-circuited stub is shunt loaded on a transmission line with two ports, transmission zeros will be created at those frequency points with $Y_{in} = \infty$, i.e., the

Fig. 3.20. Graphical solutions for transmission zeros of the two-section open-circuited stub.

Fig. 3.21. Schematic layout of the T-slot MMR stub loaded UWB bandpass filters. All units are in mm. Substrate: Roger's RT/Duriod 6010, $\epsilon_r = 10.8$ and $h = 1.27$ mm.

denominator of (3.2) equals zero. To make the analysis simple, it is assumed that the two transmission-line sections have the same electrical length, *i.e.*, $\theta_1 = \theta_2$. Hence, the condition of generating transmission zeros is written as,

$$\tan^2 \theta = \frac{Y_1}{Y_2} \quad (3.3)$$

The solutions of this equation can be found graphically by considering the expressions on the left and right hand sides as two functions, and the solutions are the intersections of these two curves as depicted in Fig. 3.20. The tangent square function on the left hand side is a periodic function with period of π , while the function on the right hand side is just a straight line along the horizontal direction. The position of the line depends on the selected characteristic admittance ratio of the two sections, *i.e.* Y_1/Y_2 . If the impedances are the same, the solutions of (3.3) are as following.

$$\theta = \frac{(2n-1)\pi}{4} \quad (3.4)$$

where n is an integer. It implies that the frequency ratio of the second transmission zero over the first transmission zero is three. As the admittance ratio changes to be greater or

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less than unity, the frequency ratios for the first two transmission zeros are able to be set larger or less than three as depicted in Fig. 3.20. In order to realize two transmission zeros at the lower/upper passband edges of the T-slot UWB bandpass filter shown in last Section, the admittance ratio required should be around 0.3. Therefore, the two solutions of (3.3) are 28.7° and 151.3° . If the first transmission zero is located around 2.3 GHz, the second transmission zero will be at 12.1 GHz, which are at two passband edges of the desired UWB passband, *i.e.*, 3.1-10.6 GHz.

To reduce the radiation loss, the characteristic impedance Z_1 is determined as the narrowest width that can be fabricated with ensured accuracy. The admittance ratio of the

(a)

(b)

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(c)

Fig. 3.22. T-slot MMR stub-loaded UWB bandpass filter. (a) Photograph. (b) Simulated and measured S-parameters. (c) Simulated and measured group delay.

stepped-impedance open-circuited stub in Fig. 3.19 is 0.29. The lengths of the two sections are around 28.7° at 2.3 GHz. It is noticed that the dimension of the T-slot MMR in the middle is unchanged as compared with the one in Fig. 3.17(a). However, instead of strong interdigital line feeding at two ports, here only relatively weak parallel-coupled lines are used. The coupling requirement is relaxed by the loading stubs. The filter depicted in Fig. 3.21 is fabricated and measured. The photograph of the fabricated circuit is shown in Fig. 3.22(a) and the measurement results are plotted in Fig. 3.22(b) and (c), respectively.

The measured frequency responses show a good UWB filtering behavior with observable transmission zeros appearing at 2.20 GHz, 2.52 GHz and 11.26 GHz. In the lower passband edge, one of the two transmission zeros is expectedly generated by the introduced loaded stubs, and the other may be created by the cross-coupling between two adjacent stubs. From Fig. 3.20, it is noticed that once the selected admittance ratio is smaller than unity, the transmission zero from next period will be moved forward. It can be utilized to suppress the higher harmonics and widen the upper stopband. The measured

results exhibit a bandpass filter response from 2.99 GHz to 10.7 GHz, *i.e.*, 112.4% fractional bandwidth at 6.85 GHz. The mid-band insertion loss is 1.4 dB and the return loss within the main passband is larger than 9.5 dB. The group delay at the passband edges become large as the trade-off of enhanced filter selectivity, and maximum group delay variation within the passband is around 1 ns. From Fig. 3.22(b), it is also noticed that there is a large amount of power loss at 16 GHz. To identify this loss, the loss factor, $(1 - |S_{11}|^2 - |S_{21}|^2)$, is calculated under two assumed cases, *i.e.*, Case I: non-perfect conductor (copper) and lossy dielectric substrate (loss tangent: 0.0023); Case II: perfect conductor and lossless dielectric substrate. The loss factor in Case I is contributed by all the losses inclusive of conductor loss, substrate loss and radiation loss. In Case II, there are no conductor and dielectric losses so that the loss is only brought by unexpected radiation. Fig. 3.23 shows the calculated loss factors for the Case I and II. Since these two factors are almost identical in high frequencies, the loss appearing around 16 GHz is primarily caused by unexpected radiation in electrically thick substrate.

3.4 Conclusion

In this Chapter, the resonance behaviors of the stub-/slot-loaded MMR are carefully studied. By loading stubs and slot at proper positions along the non-uniform transmission-line resonators, the resonant frequencies can be moved to the desired places. Together with the external coupling at input and output ports, the ultra-wideband bandpass filters can be constructed with controllable bandwidths. To enhance the filter selectivity, multi-stage MMR composed of two or more single MMRs can be used. The mode-graph with characteristics of the splitting resonant modes is studied and utilized to

Fig. 3.23. Simulated total loss factor and radiation loss factor versus frequency.

design filters with two stages. The alternative way for sharpening the roll-offs is to introduce transmission zeros at lower/upper passband edges. They can be implemented with stepped-impedance stubs loaded at two ports. In short, UWB bandpass filters are successfully constructed with stub-/slot-loaded MMRs and their performance is further improved by different approaches.

CHAPTER 4

UWB Bandpass Filters with Hybrid Microstrip/Slotline Structures

4.1 Introduction

As mentioned before, there are a lot of UWB bandpass filters that have been reportedly realized in the recent years. However, since the ultra-wide bandwidth requires very strong coupling at two ports, the gap between two parallel-coupled lines must be very small if the filter is implemented on microstrip line. The fabrication limit cannot ensure the accuracy for gaps with too small dimensions. To solve this problem, many filters were formed on two layers and make use of the wide face-to-face coupling to construct the ultra-wide passband [63]-[66], [91]-[93].

In this Chapter, the well-known back-to-back microstrip-slotline cross-junction transition [16] is rearranged and used to make up a novel class of UWB bandpass filter with lenient tolerance in fabrication. At first, a single microstrip-slotline transition is properly formed by setting the extended portions of the microstrip and slotline as one quarter of a wavelength at the central frequency of the concerned UWB passband. This transition replaces the parallel-coupled microstrip lines and/or coplanar-waveguide lines in [57]-[60] as an alternative distributed coupling structure. Then, two single transitions are directly connected through a half-wavelength uniform slotline such that a back-to-back transition based UWB bandpass filter is constructed. After an ultra-wide passband is achieved via triple resonant modes of a slotline resonator, the slotline resonator is twisted to a W or bowl shape. With this geometrical variation, one additional cross-coupling route is introduced and two transmission zeros are emerged to improve the filter selectivity. Later on, the uniform slotline resonator is replaced with stepped-impedance resonator to design UWB bandpass filters with designable bandwidth.

4.2 UWB Bandpass Filters with Slotline MMRs

4.2.1 Implementation with single uniform slotline MMR

Fig. 4.1(a) shows the basic structure of the proposed UWB bandpass filter. It is composed of two single microstrip-slotline transitions cascaded via a half-wavelength $\lambda_s/2$ uniform slotline, where λ_s is the guided wavelength of slotline at the central frequency of about 7.10 GHz. The slotline portion is etched on one side of the substrate and it is crossed at a right angle with two microstrip feed lines at about one quarter-wavelength ($l_s \approx \lambda_s/4$) from both short-circuited ends. The microstrip lines on the opposite

Fig. 4.1. (a) Geometry and (b) simulated frequency response of the proposed microstrip-slotline UWB bandpass filter with varied stub length (l_m) and fixed slotline resonator dimensions (length: 20.70mm, width:0.39 mm).

side are also extended about $l_m \approx \lambda_m/4$ (λ_m : guided wavelength of microstrip line) beyond the cross-junction positions. In order to simplify the design, uniform $50 \, \Omega$ microstrip lines are used and chamfered bends are employed to compensate the excess capacitance [16]. Fig. 4.1(b) plots the simulated frequency response of S_{21} -magnitude in a frequency range of 1.0 to 13.0 GHz for three different l_m values under the fixed $l_s \approx \lambda_s/4$. In the weak coupling situation ($l_m = 0.1$ mm), three resonant peaks can be clearly observed at 3.59 GHz, 6.96 GHz, and 10.37 GHz, respectively. They are in general contributed by the first

triple resonant modes of the slotline resonator. As l_m is gradually increased from 0.1 to 0.5 and 3.8 mm (close to one quarter-wavelength), the S_{21} -magnitude curve moves up and eventually becomes a flat roof close to 0-dB over the frequency range of 2.92 to 11.24 GHz. It implies that such an ultra-wide passband of >110% fractional bandwidth is created due to synchronous occurrence of triple resonant modes in the slotline resonator.

In order to enhance the frequency selectivity and miniaturize the overall size, a modified UWB bandpass filter is then constructed. As depicted in Fig. 4.2, the slotline

Fig. 4.2. Geometry of an UWB filter with W-shaped slotline resonator. Substrate: Roger’s RT/Duriod 6010, $\epsilon_r = 10.8$ and $h = 0.635$ mm.

resonator is twisted as a W shape and two microstrip feed lines are partially spaced in proximity. This extra coupling path may create one or two transmission zeros outside the desired UWB passband if it has the equal S_{21} -magnitude and out-of-phase with the primary transition-slotline-transition path in certain frequency points outside the UWB passband. As indicated in Fig. 4.2, the gap spacing and length of the parallel coupled-line section are indicated by s and t , respectively. As a result, a few transmission zeros can be produced outside the desired UWB passband by varying s and t due to this cross coupling behavior as exhibited in [48]-[49].

(a)

(b)

Fig. 4.3. Simulated frequency responses versus (a) Spacing (s). (b) Length (t).

Fig. 4.3(a) and (b) plot two different sets of simulated frequency responses of S_{21} -magnitude for the constituted UWB filter in Fig. 4.2. With comparison to Fig. 4.1(b), the two transmission zeros are visibly generated in the lower-stopband, facilitating the rejection skirt to be sharpened. Under the fixed coupling length of $t = 1.1$ mm, the second zero is moved up to the low cut-off edge as the spacing s is narrowed from 1.33 to 0.21

(a)

(b)

(c)

Fig. 4.4. (a) Photos of the fabricated W-shaped UWB filters with/without LPFs. (b) and (c) Measured frequency responses and group delays. LPF: low-pass filter.

mm. Meanwhile, the transmission zero near the upper passband edge varies with respect to this spacing. Under the fixed $s = 0.77$ mm, the transmission zeros in both lower and upper stopbands almost stay still as t is changed between 0.6 and 1.6 mm. However, the in-band filtering performance can be certainly improved and the upper-stopband can be slightly widened when $t = 0.6$ mm is selected. To further raise the upper stopband beyond 20 GHz, two identical open-circuited stub low-pass filters are formed along the two feed lines as discussed below.

Fig. 4.4(a) depicts the top- and bottom-view photographs of the two fabricated UWB bandpass filters, in which the dashed rectangular boxes indicate the two low-pass filter blocks. These two modified UWB filters with/without low-pass filters are designed and fabricated on the Roger’s RT/Duriod 6010 with relative permittivity: $\epsilon_r = 10.8$ and thickness: $h = 0.635$ mm. Their filtering performances are simulated and optimized using the Agilent Momentum software, whereas the fabricated filters are measured using Agilent network analyzer with the help of universal test fixture. Predicted results are well

confirmed in experiment. Fig. 4.4(b) and (c) show the measured frequency responses as well as group delays of these two UWB filters, respectively.

For the UWB filter without low-pass filters, the 3-dB fractional bandwidth is 123% at the center frequency of 6.71 GHz or its passband ranges from 2.58 to 10.83 GHz. The mid-band insertion loss is 0.97 dB, return loss is higher than 10.0 dB within the whole passband and the group delay is less than 1.00 ns with a maximum variation of 0.51 ns. For the UWB filter with low-pass filters, its filtering performances are quite similar to the one without low-pass filters in the low stopband and middle passband. The 3-dB fractional bandwidth is about 124% at the central frequency of 6.73 GHz or its passband is restricted between 2.56 and 10.90 GHz. The mid-band insertion loss is 1.05 dB, return loss is higher than 12.04 dB over the entire passband and the group delay is less than 1.07 ns with a maximum variation of 0.58 ns. However, it can be clearly seen from Fig. 4.4(b) as the low-pass filters are installed, the spurious harmonic passband around 15.2 GHz has been fully suppressed. As a result, the upper-stopband is now stretched beyond 20.0 GHz.

4.2.2 Implementation with single non-uniform slotline MMR

Fig. 4.5(a) shows the initial geometrical layout of the slotline MMR. It consists of a half wavelength high-impedance (Z_2) section in the middle and the two quarter wavelength low-impedance (Z_1) sections at two sides. The resonant frequencies of this MMR can be determined under the transmission-line resonance and they can be expressed as the solutions of the following two equations,

$$\tan \theta_2 = K \cot \theta_1 \quad (\text{odd modes}) \quad (4.1)$$

$$\tan \theta_2 = -K^{-1} \tan \theta_1 \quad (\text{even modes}) \quad (4.2)$$

where $K = Z_2/Z_1$ is the impedance ratio of two distinctive slotline sections along the MMR while θ_1 and θ_2 are their corresponding electrical lengths or phases, respectively. For simplification in this study, we assume $\theta_1 \approx \theta_2$ so that the third resonant frequency over its first counterpart can be derived from (4.1) and it can be expressed as a closed-form formula as below.

$$\frac{f_3}{f_1} = \frac{\pi}{\tan^{-1} \sqrt{K}} - 1 \tag{4.3}$$

In order to equally allocate the first three resonances close to the lower-end, center and higher-end of the targeted UWB passband, the frequency ratio f_3/f_1 should be selected slightly smaller than 3. It can be realized by setting the impedance ratio K slightly larger

(a)

(b)

Fig. 4.5. (a) Schematic layout. (b) Resonant frequency ratios versus the width ratio of the slotline MMR.

than 1 following the discussion in [118]. Fig. 4.5(b) plots the two frequency ratios, *i.e.*, f_2/f_1 and f_3/f_1 , versus the width ratio (W_2/W_1). In the design, the width (W_1) is selected as its minimum tolerable value of 0.1 mm in the existing fabrication procedure so as to minimize the unexpected conductor loss, whereas the width (W_2) is gradually raised from 0.1 to 1.0 mm. It can be observed that the first three resonances are moving closer to each other as W_2/W_1 is enlarged. When this width ratio (W_2/W_1) is chosen as 4.4, the first three resonant frequencies appear at 4.01, 7.10 and 9.84 GHz, respectively. As it will be demonstrated later on, if two microstrip-to-slotline transitions are properly introduced at the input and output ports, a dominant passband covering the FCC-defined UWB can be

(a)

(b)

Fig. 4.6. Schematic layouts of the proposed UWB bandpass filters. (a) Basic structure. (b) W-shaped structure. All units are in mm. Substrate: Roger's RT/Duriod 6010, $\epsilon_r = 10.8$ and $h = 0.635$ mm.

eventually constructed.

With similar structure as that shown in Fig. 4.2, a W-shaped UWB bandpass filter with slotline MMR is constructed and its schematic layout is depicted with its basic structure in Fig. 4.6 with all dimensions labeled. Again, the gap spacing (s) and coupled length (l) in the parallel section are adjusted to control the magnitude and phase of the cross coupling between two open-circuited stubs. Instead of defected-ground low-pass filters cascaded at input and output ports, two low-pass filters implemented on microstrip-line are used this time. As investigated previously, the transmission zeros can be properly generated at the frequency points where the transmission coefficients along the main filtering path and such a cross coupling path are equal in magnitude and 180° difference in phase. Fig. 4.7 illustrates the simulated frequency responses of the basic and modified UWB bandpass filters shown in Fig. 4.6(a) and (b). It can be clearly demonstrated herein that two transmission zeros simultaneously appear at 1.26 and 11.34

Fig. 4.7. Simulated frequency responses of the proposed UWB bandpass filters in Fig. 4.6(a) and (b) with $s = 0.48$ mm and $l = 0.6$ mm.

Fig. 4.8. W-shaped UWB bandpass filter with SISR. (a) Photographs. (b) Measured S-parameters and group delay.

GHz that result in great enhancement of the filter selectivity. As it can also be seen in Fig. 4.7, the upper stopband is significantly widened up to 20 GHz with an attenuation level higher than 14.72 dB.

The modified UWB bandpass filter with non-uniform W-shaped slotline resonator is then optimally designed and fabricated on the same material, *i.e.*, Roger's RT/Duriod 6010 with relative permittivity: $\epsilon_r = 10.8$ and thickness: $h = 0.635$ mm. Fig. 4.8(a) shows

the top- and bottom-view photographs of the fabricated filter. The measurement is done by using Agilent network analyzer and universal test fixture. The measured frequency responses and group delay are plotted in Fig. 4.8(b), which exhibits an ultra-wide passband from 2.95 to 10.16 GHz or 110% fractional bandwidth at 6.56 GHz. The mid-band insertion loss is 1.27 dB and the return loss is higher than 9.47 dB within the whole passband. The in-band group delay is less than 0.98 ns with a maximum variation of 0.41 ns. The upper stopband is really stretched up to 20 GHz, over which an insertion loss or attenuation is higher than 21.03 dB.

To further reduce the overall circuitry size, the above-formed slotline MMR with straight-line configuration can also be twisted as a bowl. Therefore, a bowl-shaped UWB bandpass filter can be constructed. The schematic layout of the proposed bowl-shaped UWB filter is depicted in Fig. 4.9 with all the dimensions labeled. With this geometrical variation, two open-ends of feed lines are moved near to each other, thus bringing out a

Fig. 4.9. Geometrical layout of the Bowl-shaped UWB bandpass filter on hybrid microstrip/slotline structure ($g = 0.7$ and $t = 2.1$). All units are in mm.

cross-coupling route that may be useful to generate transmission zeros outside the desired UWB passband. In the filter design, the gap spacing (g) and coupled length (t) in the

Fig.4.10. Simulated frequency responses of the proposed UWB bandpass filters in Fig. 4.6(a) and 4.7.

coupling section between the feed lines need to be properly determined such that the magnitude and phase of this additional coupling path can be controlled as previous examples. As studied earlier, one can figure out that the transmission zeros are primarily created at the frequency points where S_{21} -parameters of the main path and the cross coupling path are equal in magnitude and 180° difference in phase. Fig. 4.10 depicts the comparison between simulated frequency responses of the two UWB bandpass filters in Fig. 4.6(a) and 4.9. This figure clearly demonstrates that the 2nd filter really has two transmission zeros at 1.55 and 11.41 GHz, respectively, thus helpfully sharpen the roll-off rejection skirts outside the UWB passband.

As shown in Fig. 4.9, the bowl-shaped UWB bandpass filter also suffers from very narrow upper-stopband and very poor rejection at high frequencies due to unexpected high-order harmonic resonances of the slotline resonator. To address this issue, two

identical defected ground structures are installed along the two feed lines, aiming at exciting a preferred stopband beyond the desired UWB passband and eventually suppressing those unwanted harmonic resonances at high frequencies (10.6 to 20.0 GHz). After optimization designs are carried out via Agilent ADS simulator, two bowl-shaped UWB bandpass filters with and without defected ground structures are fabricated on the Roger's RT/Duriod 6010. Fig. 4.11(a) illustrates the top- and bottom-view photographs of these two filters. The defected ground structure is composed of three transverse slots that are formed on the ground plane underneath the feed-line strip conductor. As explained above, these two defected ground structures are used to produce a highly attenuated stopband beyond the desired UWB passband.

The two fabricated filters are then measured with the help of Agilent network analyzer and universal test fixture. Fig. 4.11(b) and (c) plot the measured frequency responses of scattering parameters and group delays, respectively. For the UWB bandpass filter without defected ground structures, the measured results have clearly shown the existence of a wide passband from 3.23 to 10.60 GHz or 107% fractional bandwidth at the central frequency of 6.92 GHz. The results also show us that the mid-

(a)

Fig. 4.11. (a) Photographs of the fabricated UWB filters with/without DGSs. (b) and (c) Measured scattering parameters and group delays. DGS: defected ground structure.

band insertion loss achieves 1.46 dB, the return loss within the whole passband is higher than 12.75 dB and the group delay is less than 0.86 ns with a maximum variation of 0.37 ns.

On the other hand, for the UWB bandpass filter with defected ground structures, the upper stopband has been stretched up to 18.2 GHz with the insertion loss higher than 16.0

dB over the upper stopband up to 20 GHz. But, due to the extra effects of the defected ground structures, the filtering performance in the upper side of such a passband has been unfortunately degraded a little. In this case, the measured passband ranges from 3.25 to 10.21 GHz or 103% FBW at the central frequency of 6.73 GHz. The mid-band insertion loss is now about 1.55 dB, the in-band return loss is still higher than 10.43 dB and the group delay is less than 0.84 ns with a maximum variation of 0.34 ns.

4.2.3 Implementation with three uniform/non-uniform slotline/microstrip line MMRs

In the previous Sections, the UWB bandpass filters were formed making use of the

Fig. 4.12. Schematic layout of the UWB bandpass filter with cascaded uniform/non-uniform resonators ($g = 1.64$ and $t = 1.58$). All units are in mm.

back-to-back microstrip line to slotline transitions and the single multiple-model slotline resonators. The filter selectivity is enhanced by introducing the cross-coupling route between the input and output ports, and the upper stopband is widened with the low-pass filters on the feed lines. In this Section, an alternative way to design an UWB bandpass

Fig. 4.13. Simulated frequency responses of the seven-mode resonator under weak/strong input/output coupling.

filter with shape rejection skirts and wide upper stopband is introduced by using three mixed MMRs. Two uniform slotline resonators and one non-uniform microstrip line resonator are cascaded through microstrip-to-slotline transitions and are considered as one resonator. The first two resonant modes of the slotline resonator and the first three resonances of the microstrip line resonator are used to construct the UWB passband. The dimensions of the resonators used are determined such that the seven resonances are located uniformly within the targeted passband. With the microstrip-to-slotline coupling at two ports, ultra-wide passband is established.

The schematic layout of this UWB filter is depicted in Fig. 4.12. As shown in the figure, the two microstrip feed lines are arranged such that the two open-circuited ends are close to each other to bring in the cross-coupling route as the previous examples. The gap g and facing length l are also used to adjust the positions of the two transmission zeros at the lower/higher passband edges. Fig. 4.13 shows the simulated frequency

responses of the cascaded seven-mode resonator under weak and strong coupling. As observed from the figure, seven resonances are clearly seen and an UWB filtering behavior with high selectivity and wide upper stopband are also noticed.

4.3 Conclusion

In this Chapter, a class of UWB filters with slotline resonator and microstrip-to-slotline transitions is proposed and studied. The characteristic of slotline resonator resonance behavior is analyzed. The basic filter structure is composed of a single uniform or non-uniform slotline resonator and microstrip-to-slotline transitions as distributive coupling structures. By adjusting the impedance ratio of different sections of the slotline resonator, the UWB filters with controllable bandwidth can be constructed. To reduce the total circuitry size and introduce a cross-coupling path between the input and output ports, the slotline resonator is twisted to W or bowl shape. With such geometrical variation, the filter size is miniaturized and two transmission zeros are generated at lower/upper passband edges to enhance the filter selectivity. The upper stopband performance can be significantly improved by installing low-pass structures along the feed lines. Later on, a multi-stage UWB filter with both microstrip and slotline resonator is studied and designed and it shows good filtering response with sharp rejection skirts.

CHAPTER 5

Synthesis Design of UWB Bandpass Filter with Hybrid Transmission-Line Structures

5.1 Introduction

As discussed earlier, the spectrum masks released by the FCC that are depicted in Fig. 1.2(a) and (b) are very strict. Therefore, the UWB bandpass filters used to regulate the UWB signals have to be designed with very high selectivity or very sharp rejection skirts at the passband edges. It could be achieved by either using the conventional high-pass prototype filters with high order [113] or purposely introducing two transmission

zeros at the lower-/higher-passband edges [48]-[49], or both of these two methods. However, the prototype filter with higher order will result in a large circuitry area and the additional cross-coupled route between input and output port requires a lot of simulation effort to achieve the desired performance.

Let us take a look at the insertion loss function of the conventional high-pass prototype or the proposed high-pass transmission-line model introduced in Chapter 2 with series open-circuited stubs, and it is rewritten in the following,

$$P_L = 1 + \left| \frac{P_{2n-1}(\sin \theta)}{\cos \theta} \right|^2 \quad (5.1)$$

It is observed that the denominator of the characteristic function within the absolute value function is $\cos \theta$. Therefore, we only have 1st – order transmission zeros at multiple integer times of π ($0, \pi, 2\pi, \dots$). If the order of denominator can be increased, the order of transmission zeros will be increased as well. As a result, shaper rejection skirts can be established.

In the high-pass prototypes shown in Fig. 2.2(b) and Fig. 2.3(b), only one type of stubs is involved, either shunt short-circuited stubs or series open-circuited stubs. It gives the inspiration of using composite stubs that consist of both shunt and series stubs. In [111], the generating functions with high-pass or quasi-bandpass responses were given, and from the theoretical examples, it may be figured out that the higher order transmission zeros could be introduced by the hybrid stubs. In this Chapter, a general transmission-line model for a class of quasi-bandpass filters with designable number of in-band transmission poles is proposed and analyzed with synthesis approach. The network consists of mixed stubs connected through uniform transmission lines. After determining all the involved element values by synthesis method, the transmission-line

Fig. 5.1. Transmission-line model of the proposed single stage band-pass filter prototype based on composite stubs and an i -section non-uniform connecting line. (a) i is odd-integer (1, 3, 5...). (b) i is even-integer (0, 2, 4, 6...).

model is realized using hybrid microstrip line and slotline to verify the theoretical predictions. Later on, a multi-stage UWB filter transmission-line model is proposed and analyzed with similar procedure.

5.2 General Transmission-Line Model of UWB Bandpass Filter with Composite Stubs

5.2.1 General single-stage ($N = 1$) transmission-line model

The proposed general transmission-line model is composed of two sets of series open-ended and shunt short-ended stubs that are connected through an i -section non-uniform transmission line in the middle. It is depicted in Fig 5.1(a) and (b) for the odd- and even-integer i , respectively. All the series and shunt stubs are set with the same

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electrical length, θ_c , at lower cutoff frequency, f_c . The normalized impedances for the series and shunt stubs are labeled as z_{om} and z_{sm} , where m indicates the m -th set of composite stubs. Meanwhile, each section of the middle connecting line is exactly chosen as θ_c , and the normalized impedances of the first section to the one at the center are named from $z_{m,m+1}^1$ to $z_{m,m+1}^{(i+1)/2}$ for the odd-integer i (1, 3, 5, ...) and to $z_{m,m+1}^{i/2}$ for the even-integer i (0, 2, 4, 6, ...). The subscript indicates the connecting line is between the m -th and $(m + 1)$ -th sets of stubs, and the superscript indicates the section number. Plus, the network is horizontally symmetrical while the input and output are terminated by two unit normalized impedances, *i.e.*, $z_0 = 1$. The overall $ABCD$ matrix of the proposed transmission-line model in Fig. 5.1 can be obtained by multiplying the $ABCD$ matrices of individual sections in sequence. Hence, the squared magnitude of transmission coefficient S_{21} can be expressed in terms of a characteristic function F as

$$|S_{21}|^2 = \frac{1}{1 + |F|^2} \quad (5.2)$$

and

$$F = \frac{B - C}{2} \quad (5.3)$$

where B and C are the entries of the overall $ABCD$ matrix of the transmission-line network in Fig. 5.1. By substituting B and C and manipulating (5.3), the characteristic function F can be simplified as

$$F = \frac{P_{i+3}(\cos \theta)}{\sin^3 \theta} \quad (5.4)$$

where P_{i+3} is an even polynomial with highest order $(i + 3)$. It is noticed that the highest order of the denominator of F becomes three, which means the transmission zero

emergence at multiple integer times of π are 3rd – order transmission zeros. In order to regulate the frequency response to exhibit Chebyshev equal-ripple characteristics, the targeted function must be chosen in the format of (5.5) as discussed in [109], [111]. It results in a Chebyshev equal-ripple response with $(n + q)$ in-band transmission poles and generates q^{th} -order transmission zeros at $x = \pm\alpha$.

$$|S_{21}|^2 = \frac{1}{1 + \varepsilon^2 \cos^2(n\phi + q\xi)} \quad (5.5)$$

and

$$x = \cos\phi = \alpha \cos\theta \quad (5.6a)$$

$$y = \cos\xi = x \sqrt{\frac{\alpha^2 - 1}{\alpha^2 - x^2}} \quad (5.6b)$$

where ε is the specified equal-ripple constant in the concerned passband and it relates to the ripple level L_A as,

$$\varepsilon = \sqrt{10^{\frac{L_A}{10}} - 1} \quad (5.7)$$

$\alpha = 1/\cos\theta_c$, where θ_c is the phase at the lower cutoff frequency. Due to the frequency-distribution characteristic of transmission lines, the upper cutoff frequency of this band-pass filter appears at $(180^\circ - \theta_c)$. Based on the relationship among those variables in (5.6a) and (5.6b), the cosine function in (5.5) can be expanded as

$$\cos(n\phi + q\xi) = T_n(x)T_q(y) - U_n(x)U_q(y) \quad (5.8)$$

where $T_n(x)$ and $U_n(x)$ are the Chebyshev polynomial functions of the first and second kinds of degree n , respectively. Using the identity for $T_n(x)$ and $U_n(x)$

$$T_{n+1}(x) - xT_n(x) = -\sqrt{1 - x^2}U_n(x) \quad (5.9)$$

the targeted transfer function in (5.5) can be analytically expressed in terms of these two kinds of Chebyshev functions. To equalize (5.2) with (5.5), n and q in (5.5) should be selected as i and 3. There will be $(i + 4)/2$ or $(i + 5)/2$ sets of equations derived depending on whether i is even or odd. After setting the specified ripple constant and cutoff phase,

(a)

(b)

Fig. 5.2. Comparison among the proposed multi-pole bandpass filters with different i (ripple-level, $L_A = 0.1$ dB). (a) Improved selectivity with the same $\theta_c = 45^\circ$. (b) Extendable bandwidth with the same rejection skirts.

the unknown element values in the network can be determined by solving the equations.

Fig. 5.2(a) depicts the comparison in $|S_{21}|$ of the prototype filter with different i . It is observed herein that the number of in-band transmission poles is exactly equal to $(i + 3)$, while the ripple level and cutoff phase are 0.1 dB and 45° as predetermined. The extra transmission poles are introduced by the additional section in the connecting line. By comparing these four sets of frequency responses, it is noticed that the ones with more transmission poles in the passband offer sharper rejection skirts under the same ripple level and cutoff phase. The proposed filters also provide higher selectivity as compared with the conventional filters with pure shunt short stubs [46]-[49] and series open stubs [119]-[122] if they are designed under the same specifications, *i.e.*, ripple level, cutoff phase and number of in-band poles. The highest order in the denominator of F here is 3, therefore, 3rd-order transmission zeros are generated at $\theta = 0^\circ$ and 180° , while the ones in [46]-[49], [119]-[122] are all 1st-order zeros. This issue will be illustrated in the following sections in detail. Fig. 5.2(b) is plotted to demonstrate that more transmission poles can be applied to further widen the bandwidth of the core passband at no cost of extra resonators. As shown in Fig. 5.2(b), the fractional bandwidth with the same selectivity is greatly extended from 103.9% to 142.0%, when two extra poles are introduced into passband as i increases from 1 to 3.

5.2.2 Synthesis procedure for UWB bandpass filter

If there is only one transmission-line section between two sets of hybrid stubs in the transmission-line model shown in Fig. 5.1(a), *i.e.*, $i = 1$, the element values involved in this transmission-model are z_{o1} , z_{s1} and $z_{1,2}^1$, since the transmission-line model is

symmetric with respect to the central plane. Following the synthesis procedure mentioned above, the characteristic function F can be derived from the transmission-line network as,

$$F = j(k_1 \frac{\cos^4 \theta}{\sin^3 \theta} + k_2 \frac{\cos^2 \theta}{\sin^3 \theta} + k_3 \frac{1}{\sin^3 \theta}) \quad (5.10)$$

where

$$k_1 = \frac{z_{o1}^2 z_{1,2}^1 - z_{1,2}^1}{2z_{s1}^2} + \frac{z_{o1}^2 + z_{o1} z_{1,2}^1 - 1}{z_{s1}} + \frac{z_{o1}^2 - 1}{2z_{1,2}^1} + z_{o1} + \frac{z_{1,2}^1}{2} \quad (5.11a)$$

$$k_2 = -\frac{z_{o1}^2 - 2}{2z_{1,2}^1} - \frac{z_{o1} z_{1,2}^1 - 1}{z_{s1}} - z_{o1} - z_{1,2}^1 + \frac{z_{1,2}^1}{2z_{s1}^2} \quad (5.11b)$$

$$k_3 = \frac{z_{1,2}^1}{2} - \frac{1}{2z_{1,2}^1} \quad (5.11c)$$

In order to make the derived insertion loss function equal to the targeted one in (5.5), the n and q in (5.8) have to be selected as 1 and 3. Therefore, the characteristic function in the denominator of (5.5) can be expanded by substituting $n = 1$ and $q = 3$ in (5.8) as the following,

$$\cos(\phi + 3\xi) = \cos\phi \cos 3\xi - \sin\phi \sin 3\xi \quad (5.12)$$

and

$$\cos\phi = T_1(x) = \alpha \cos\theta \quad (5.13a)$$

$$\sin\phi = U_1(x) = -\frac{1}{\sqrt{1-x^2}}(T_2(x) - xT_1(x)) \quad (5.13b)$$

$$\cos 3\xi = T_3(y) = 4y^3 - 3y \quad (5.13c)$$

$$\sin 3\xi = U_3(y) = -\frac{1}{\sqrt{1-y^2}}(4y^4 - 5y^2 + 1) \quad (5.13d)$$

By substituting (5.13) into (5.12), the following can be derived,

$$\cos(\phi + 3\xi) = x(4y^3 - 3y) - \frac{1}{\sqrt{(1-x^2)(1-y^2)}}(x^2 - 1)(4y^4 - 5y^2 + 1) \quad (5.14)$$

Replacing y in terms of x as (5.6b), the cosine function in the denominator of (5.5) can be simplified as,

$$\begin{aligned} \cos(\phi + 3\xi) = & (4\alpha(\alpha^2 - 1)^{\frac{3}{2}} + 4\alpha^4 + 3\alpha\sqrt{\alpha^2 - 1} - 3\alpha^2) \frac{\cos^4 \theta}{\sin^3 \theta} \\ & + (-3\alpha\sqrt{\alpha^2 - 1} - 5\alpha^2 + 3) \frac{\cos^2 \theta}{\sin^3 \theta} + \frac{1}{\sin^3 \theta} \end{aligned} \quad (5.15)$$

After equalizing k_1 , k_2 , and k_3 in (5.10) with the coefficients of $\cos^4 \theta / \sin^3 \theta$, $\cos^2 \theta / \sin^3 \theta$ and $1 / \sin^3 \theta$ in (5.15) and multiplying by the ripple constant ε , the following set of equations can be derived.

$$\varepsilon(4\alpha(\alpha^2 - 1)^{\frac{3}{2}} + 4\alpha^4 + 3\alpha\sqrt{\alpha^2 - 1} - 3\alpha^2) = k_1 \quad (5.16a)$$

$$\varepsilon(-3\alpha\sqrt{\alpha^2 - 1} - 5\alpha^2 + 3) = k_2 \quad (5.16b)$$

$$\varepsilon = k_3 \quad (5.16c)$$

Therefore, the normalized characteristic impedances, z_{o1} , z_{s1} , and $z_{1,2}^1$, for the series open-ended stub, shunt short-ended stub and the middle connecting line can be explicitly determined once ε and θ_c or fractional bandwidth $(180^\circ - \theta_c) / \theta_c$ are specified. Particularly, (5.16c) indicates that the normalized characteristic impedance of the connecting line with the length of θ_c only relies on the specified ripple constant ε .

Using the same analysis approach, the characteristic function of the filter prototype shown in Fig. 5.1(b) with $i=2$ and stage number $N = 1$ can be derived as

$$F = j(k_1 \frac{\cos^5 \theta}{\sin^3 \theta} + k_2 \frac{\cos^3 \theta}{\sin^3 \theta} + k_3 \frac{\cos \theta}{\sin^3 \theta}) \quad (5.17)$$

where

$$k_1 = \frac{z_{o1}^2 z_{1,2}^1 - z_{1,2}^1}{z_{s1}^2} + \frac{2(z_{o1}^2 + z_{o1} z_{1,2}^1 - 1)}{z_{s1}} + \frac{z_{o1}^2 - 1}{z_{1,2}^1} + 2z_{o1} + z_{1,2}^1 - \frac{1}{z_{1,2}^1} \quad (5.18a)$$

$$k_2 = -\frac{z_{o1}^2 - 2}{z_{1,2}^1} - \frac{z_{o1}^2 + 2z_{o1} z_{1,2}^1 - 3}{z_{s1}} - 3z_{o1} - 2z_{1,2}^1 + \frac{z_{1,2}^1}{z_{s1}^2} \quad (5.18b)$$

$$k_3 = z_{o1} + z_{1,2}^1 - \frac{1}{z_{s1}} - \frac{1}{z_{1,2}^1} \quad (5.18c)$$

Looking at (5.17) and (5.10) comparatively, the highest order in denominator is unchanged, but that in numerator is increased by 1. To achieve the equalization of (5.17) with the characteristic function in (5.5), n and q in (5.8) now have to be chosen as 2 and 3, respectively. As a result, a complicated but explicit expression can be derived as given in (5.19) and it represents the necessary condition in this synthesis procedure.

$$\begin{aligned} \cos(2\phi + 3\xi) = & (8\alpha^2(\alpha^2 - 1)^{\frac{3}{2}} + 8\alpha^5 + 6\alpha^2\sqrt{\alpha^2 - 1} - 6\alpha^3) \frac{\cos^5 \theta}{\sin^3 \theta} \\ & + (-6\alpha^2\sqrt{\alpha^2 - 1} - 4(\alpha^2 - 1)^{\frac{3}{2}} - 10\alpha^3 - 3\sqrt{\alpha^2 - 1} + 6\alpha) \frac{\cos^3 \theta}{\sin^3 \theta} \\ & + (3\sqrt{\alpha^2 - 1} + 2\alpha) \frac{\cos \theta}{\sin^3 \theta} \end{aligned} \quad (5.19)$$

By equalizing the corresponding coefficients for different-order variables in (5.17) and (5.19), a set of equations can be derived as (5.20a) to (5.20c) and they are similar to those in (5.16a) to (5.16c). By solving these equations, normalized characteristic impedances for all the involved stubs and connecting line can be calculated.

Fig. 5.3. Predicted frequency responses of two bandpass filter prototypes under varied electrical length θ_c . (a) $i=1$ (b) $i=2$.

$$\varepsilon(8\alpha^2(\alpha^2-1)^{\frac{3}{2}}+8\alpha^5+6\alpha^2\sqrt{\alpha^2-1}-6\alpha^3)=k_1 \quad (5.20a)$$

$$\varepsilon(-6\alpha^2\sqrt{\alpha^2-1}-4(\alpha^2-1)^{\frac{3}{2}}-10\alpha^3-3\sqrt{\alpha^2-1}+6\alpha)=k_2 \quad (5.20b)$$

$$\varepsilon(3\sqrt{\alpha^2-1}+2\alpha)=k_3 \quad (5.20c)$$

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As the element values are determined according to the specified ripple constant (ϵ) and lower cutoff frequency (θ_c), all the coefficients in the characteristic functions (5.10) and (5.17) can be calculated via (5.16a)-(5.16c) and (5.18a)-(5.18c). As such, $|S_{21}|^2$ can be derived from (5.10). Fig. 5.3(a) and (b) plot the frequency responses of two single stage Chebyshev band-pass prototypes with $i = 1$ and 2, $\theta_c = 30^\circ, 45^\circ$ and 60° and fixed ripple level $L_A = 0.1$ dB. As shown in the inserted frames of these figures, the upper cutoff phase ($180^\circ - \theta_c$) is exactly equal to $150^\circ, 135^\circ$ and 120° as its lower counterpart (θ_c) is selected as the above three values. On the other hand, the ripple level is all restricted to 0.1 dB as pre-specified. Moreover, the simulated results show that the passbands repeatedly appear at the periodicity of 180° . Thus, the next passband will start at $(180^\circ + \theta_c)$ and end at $(360^\circ - \theta_c)$. Looking at Fig. 5.3(a) and (b) comparatively, the total number of in-band transmission poles for these two filter prototypes are 4 and 5, respectively. As a result, the five-pole band-pass filter has a larger cutoff attenuation skirt than its four-pole counterpart under the same specifications, *i.e.*, ϵ and θ_c . The extra transmission pole in Fig. 5.3(b) is introduced by an extra section in the middle line. If the prototype filters have i -section non-uniform transmission line as shown in Fig. 5.1, the highest degree of the numerator for the derived characteristic function F will equal $(i + 3)$. It implies that the number of transmission poles within the passband is $(i + 3)$ which is the number of solutions that make $|F|$ equal zero. On the other hand, the highest degree in the numerator while expanding (5.8) is $(n + q)$. Therefore, in order to equalize the magnitude of two characteristic functions, q and n should equal to 3 and i , respectively.

Now, let us compare the proposed single stage filters with the reported ones in [46]-[49], [119]-[122] that have either shunt short-ended, series open-ended stubs or both.

Observing the polynomial functions in (5.10) and (5.17), one can understand that the number of in-band transmission poles equals the highest order of $\cos\theta$ in the numerator. The highest order of $\sin\theta$ in denominator is $q = 3$ and it represents the existence of 3rd-order transmission zeros in the lower and upper ends of the passband, *i.e.*, $\theta = 0^\circ$ and 180° . It needs to be highlighted that there exist only 1st-order transmission zeros in the high-pass prototypes with purely shunt/series stubs. Thus, the out-of-band roll-off or attenuation skirts should be significantly enhanced, especially for the multi-stage prototype with multiplying attenuation that will be introduced later. Fig. 5.4 plots the simulated $|S_{21}|$ -graphs of the filter in [46] or [119] and the proposed filter, all with five poles in passband for comparative study. It is quantitatively confirmed that the proposed filter prototype with composite series/shunt stubs has much higher attenuation skirts in the lower stopband of $0 < \theta < \theta_c$ and the upper stopband band of $(180^\circ - \theta_c) < \theta < 180^\circ$ than those with either series or shunt stubs. In [126], a dual network of the proposed

Fig. 5.4. Comparison in $|S_{21}|$ between the conventional and proposed five-pole bandpass filter prototypes with purely shunt stubs [46] or series stubs [119] and composite stubs.

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single stage transmission-line model with $i = 0$ was utilized to design low-pass filter with out-of-band performance improvement.

Stemming from the proposed single stage filter prototypes shown in Fig. 5.1(a) and (b) for $i = 1$ and 2, two bandpass filters with four and five poles will be designed and implemented on hybrid slotline and microstrip line structures in this part. To account for the distinctive features of these two transmission lines, the shunt short-ended stubs are

(a)

(b)

Fig. 5.5. Schematic layouts of two proposed ultra-wideband bandpass filters implemented on hybrid microstrip and slotline structures. (a) Four-pole. (b) Five-pole. MSL: microstrip line.

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formed as slotline while the series open-ended stubs are constructed in microstrip line. In synthesis design based on the filter prototypes in Fig. 5.1, the microstrip-to-slotline right-angle cross-transition section can be simply expressed as an equivalent transformer with a unity turns ratio since the substrate is selected electrically thin. The substrate used in the work is RT/Duroid 6010 with a dielectric constant of 10.8 and thickness of 0.635 mm. Using the closed-form design formula for slotline and microstrip line in [16], the schematic layouts of two filters can be initially determined using the above synthesis procedure and they are depicted in Fig. 5.5(a) and (b) with all the dimensions indicated.

As illustrated in Fig. 5.5(a) and (b), the microstrip line is formed on one side of the substrate, while the slotline is etched on the opposite side that is perpendicular with the microstrip line. The electrical lengths from the microstrip-line open-ends and slotline short-ends to the intersection points are both equal to θ_c . The connecting slotline is set as θ_c or $2\theta_c$. In this design, $\theta_c = 45^\circ$ and 40° are readily selected, aiming to work out the four- and five-pole band-pass filters with ultra-wide passband. From the procedure discussed previously, three normalized impedances, z_{o1} , z_{s1} , and $z_{1,2}^1$ are at first determined such that the characteristic impedances, Z_{o1} , Z_{s1} , and $Z_{1,2}^1$ of all the stubs and connecting line in these two filters can be calculated. Table 5.1 and 5.2 tabulate the values of these impedances.

TABLE 5.1
CIRCUIT PARAMETERS FOR FOUR-POLE FILTER ($\theta_c = 45^\circ$ at 3 GHz)

Stub line impedance	Connecting line impedance
$Z_{o1} = 54.39 \, \Omega$	$Z_{1,2}^1 = 58.21 \, \Omega$
$Z_{s1} = 59.70 \, \Omega$	$Z_0 = 50 \, \Omega$

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TABLE 5.2
CIRCUIT PARAMETERS FOR FIVE-POLE FILTER ($\theta_c = 40^\circ$ at 3 GHz)

Stub line impedance	Connecting line impedance
$Z_{o1} = 46.89 \, \Omega$	$Z_{1,2}^1 = 70.48 \, \Omega$
$Z_{s1} = 58.47 \, \Omega$	$Z_0 = 50 \, \Omega$

Without any change in filter dimensions as shown in Fig. 5.5(a) and (b), full-wave simulator is employed to simulate their overall filter layouts and two relevant filter circuits are fabricated for experiment. Fig. 5.6 and 5.8 illustrate three sets of S-

(a)

(b)

Fig. 5.6. Four-pole ultra-wideband bandpass filter in Fig. 5.5(a). (a) Photographs. (b) Predicted and measured S-parameters. (c) Predicted and measured group delays.

magnitudes and groups delays which are obtained from the transmission-line model, full-wave simulation and microwave measurement, respectively. All these three sets of frequency responses and group delays are found in good agreement with each other over the plotted frequency range. The detailed comparisons of the electrical parameters are tabulated in Table 5.3 and 5.4 for these two band-pass filters. Four and five transmission poles can be clearly observed in the realized wide passband for both two filters. The positions of the poles and zeros match well with each other, especially for the two sets of graphs obtained from transmission-line model and full-wave simulation. Meanwhile, the in-band group delays for the four- and five-pole bandpass filters are all flat in the middle passband, and become higher near the lower and upper cutoff frequencies as the trade-off of high rejection skirts. Similar results can also be observed in reported works.

On the other hand, some visible discrepancies exist among the three sets of results. To our knowledge, the extra poles appearing in the measurement results are primarily brought in by multiple reflections along the feed lines due to the unexpected discontinuity effects in the SMA connectors and input/output terminals of these filters. A coaxial-to-

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microstrip transition can be replaced by a simply equivalent circuit as depicted in Fig. 5.7(a), where L_s and C_s are the series inductance and shunt capacitance, respectively [16]. The value of L_s and C_s are functions of the substrate thickness. For the material with thickness of 25mil, the shunt capacitor effect is dominant. To investigate the effect of SMA connectors on the return loss performance, the four-pole bandpass filter in Fig. 5.5(a) with different shunt capacitances at two terminals is simulated and the frequency responses are plotted in Fig. 5.7(b). From this figure, it can be observed that as the capacitance increases, there are extra poles introduced in the passband, the return loss increases and becomes unbalanced.

Fig. 5.7. (a) Equivalent circuit representation of a coaxial-to-microstrip transition. (b) S-parameter of four-pole bandpass filter in Fig. 5.5(a) with different C_s and L_s is ignored.

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Fig. 5.8. Five-pole ultra-wideband bandpass filter in Fig. 5.5(b). (a) Photograph. (b) Predicted and measured S-parameters. (c) Predicted and measured group delays.

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The frequency shift at high frequencies between the predicted and measured results may be caused by many uncontrollable factors, such as high frequency dispersion of the non-TEM slotline mode, frequency-dependent permittivity, misalignment of slot/strip positions in two layers and tolerance in etching fabrication. Radiation loss in the slotline's area can be certainly reduced by packaging these filters in a metallic enclosure, but it increases the total circuitry size and generates unexpected cavity resonances that are strongly dependent on the length/width of this enclosure. The primary purpose of implementing these filters on hybrid microstrip line and slotline structures is to give a

TABLE 5.3
COMPARISONS FOR FOUR-POLE BANDPASS PROTOTYPE FILTER

	Synthesis	EM simulation	Measurement
3 dB lower cut-off	2.57 GHz	2.64 GHz	2.59 GHz
Maximum in-band insertion loss	0.1 dB	0.95 dB	1.52 dB
Minimum in-band return loss	16.43 dB	13.16 dB	8.45 dB
3 dB upper cut-off	9.43 GHz	9.06 GHz	8.84 GHz
Maximum in-band group delay variation	0.27 ns	0.33 ns	0.31 ns

EM: electromagnetic.

TABLE 5.4
COMPARISONS FOR FIVE-POLE BANDPASS PROTOTYPE FILTER

	Synthesis	EM simulation	Measurement
3 dB lower cut-off	2.63 GHz	2.72 GHz	2.66 GHz
Maximum in-band insertion loss	0.1 dB	0.74 dB	1.55 dB
Minimum in-band return loss	16.43 dB	13.43 dB	10.15 dB
3 dB upper cut-off	10.86 GHz	10.52 GHz	9.98 GHz
Maximum in-band group delay variation	0.33 ns	0.43 ns	0.34 ns

EM: electromagnetic.

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quantitative verification on the synthesis design approach based on the proposed transmission-line model. Overall, both simulated and measured results have confirmed the feasibility and accuracy of the developed synthesis procedure very well.

It is interesting to compare the proposed prototype with those reported in [67]-[68]. In this case, the latter ones can be treated as the specified cases of the former one with $i=2$. The original idea in [67] and [68] is to consider the overall slotline section as a triple-mode resonator with uniform and non-uniform configurations. In particular, the non-uniform section with different characteristic impedances provides more flexibility and capability in the passband selection. With proper adjustment of impedance ratio, the first three resonances can be relocated uniformly within the desired passband. Together with the tight coupling of microstrip-to-slotline transitions with two additional transmission poles, an ultra-wide passband with varied bandwidth and five in-band poles can be realized. By observing the element values obtained from the above synthesis design, it can be found that as the fractional bandwidth increases, the normalized characteristic impedances become higher for two shunt short-ended stubs and lower for the middle line. This phenomenon exactly matches the multiple-mode theory as reported in [67] and [68]. On the other hand, in the above proposed prototype, the short-/open-circuited stubs have the electrical length θ_c at lower cutoff frequency f_c , that is equal to 90° at center frequency of the passband. This approach well explains and verifies the designs in [67]-[68] from the synthesis design point of view. Therefore, instead of time-consuming cut-and-try method via full-wave simulators, this synthesis method allows us to efficiently design these wideband filters using a set of closed-form design formula.

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(a)

(b)

(c)

Fig. 5.9. Comparison between measured results of two five-pole bandpass filters with and without periodical loading of three strips. (a) Photograph. (b) Predicted and measured S-parameters. (c) Predicted and measured group delays.

In order to improve the upper stopband performance of the proposed filters, the transverse strips are periodically placed above the middle connecting slotline such that the high order harmonic passbands can be effectively suppressed as done in [69]. One design example is given for the five-pole filter shown in Fig. 5.5(b). The photograph of the fabricated filter is shown in Fig. 5.9(a). The three strips are 3.30 mm in length and 0.20 mm in width, while the periodicity or spacing between them is 3.40 mm. Fig. 5.9(b) and (c) show the measured S-parameters and group delays of the two filters with and without three attached strips, respectively. Obviously, the in-band frequency responses are almost unchanged, but the upper stopband is largely pushed from 14.0 to 18.0 GHz with an attenuation of higher than 18.82 dB.

From the calculated element values for $i = 1$ and 2 with ripple constant $\varepsilon = 0.1$ dB shown in Appendix III, it is noticed as the filter order increases, bandpass filter with wider bandwidth is much easier to implement in practice with hybrid microstrip line and slotline structures. Another design example with $i = 3$ for single stage filter prototype depicted in Fig. 5.1(a) is given to illustrate the design of bandpass filters with 130%

Fig. 5.10. Schematic layout of the proposed six-pole UWB bandpass filter. All units are in mm. MSL: microstrip line.

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fractional bandwidth. The proposed transmission-line model is implemented on the same substrate with thickness of 0.635 mm and a dielectric constant of 10.8. Since the number of transmission-line sections between the two sets of stubs is three, there will be six transmission poles generated in the passband. The schematic layout of this six-pole filter is illustrated in Fig. 5.10. Again, the series open-circuited and shunt short-circuited stubs are realized by microstrip line and slotline with open or short ends on two layers, respectively. The normalized impedances for the stubs and connecting lines z_{o1} , z_{s1} , $z^1_{1,2}$, $z^2_{1,2}$ and $z^3_{1,2}$ are calculated by the procedure shown previously and the design specifications are given in Table 5.5. The references at two ports are selected as 50 Ω .

TABLE 5.5
CIRCUIT PARAMETERS FOR SIX-POLE FILTER ($\theta_c = 30^\circ$ AT 3 GHz)

Stub line impedance	Connecting line impedance
$z_{o1} = 0.65$	$z^1_{1,2} = z^3_{1,2} = 1.41$
$z_{s1} = 1.65$	$z^2_{1,2} = 1.72$

(a)

(b)

(c)

Fig. 5.11. Six-pole ultra-wideband bandpass filter. (a) Photographs. (b) Predicted and measured S-parameters. (c) Predicted and measured group delays.

After optimization via EM simulator to take all the discontinuities into account, the prototype filter is fabricated and its photograph is shown in Fig. 5.11(a). The measured frequency responses and group delay are plotted with the simulated ones in Fig. 5.11(b) and (c), respectively. Basically, the measured results agree with the predicted ones reasonably well. It shows the measured bandpass filter with 132.9% fractional bandwidth at center frequency 7.81 GHz. The mid-band insertion loss is 0.72 dB; return loss within

the main passband is larger than 10.42 dB and the maximum group delay variation is 5.13 ns. The visible frequency shifts at higher frequency is primarily brought by the unexpected dispersion in slotline and frequency dependency of the substrate as mentioned before.

5.2.2 General multiple-stage transmission-line model

To extend the above synthesis approach to the multi-stage filter design with improved out-of-band attenuation performance, a three-stage band-pass filter prototype with $i = 1$ will be studied herein and its transmission-line model is shown in Fig. 5.12.

Fig. 5.12 Transmission-line model of a three-stage ($N = 3$) band-pass filter prototype with $i = 1$.

This filter is composed of cascading four sets of composite stubs through three sections of transmission line. In the synthesis design, all the series stubs, shunt stubs and connecting lines have a fixed electrical length as θ_c at the specified lower cutoff frequency. The normalized characteristic impedances of the first and second sets of hybrid stubs are indicated as z_{o1} , z_{s1} and z_{o2} , z_{s2} , respectively, while the ones for the connecting lines between the first set and second set of stubs, the second and the third set of stubs are $z_{1,2}^I$ and $z_{2,3}^I$. Since the entire network is symmetrical with respect to the

Fig. 5.13. Schematic layout of the proposed three-stage ultra-wideband bandpass filter shown in Fig. 5.11.

central plane, the element values of the right half of the network are the same as their corresponding mirror images on the left.

Again, the $ABCD$ matrix of the whole network can be obtained easily by multiplying the individual $ABCD$ matrices of each section in sequence and the characteristic function F is derived as (5.21) and the coefficients of the polynomial are functions of the six element variables as shown in (5.22) ,

$$F = j(k_1 \frac{\cos^8 \theta}{\sin^5 \theta} + k_2 \frac{\cos^6 \theta}{\sin^5 \theta} + k_3 \frac{\cos^4 \theta}{\sin^5 \theta} + k_4 \frac{\cos^2 \theta}{\sin^5 \theta} + k_5 \frac{1}{\sin^5 \theta}) \quad (5.21)$$

where

$$k_1, k_2, k_3, k_4, k_5 = f(z_{o1}, z_{s1}, z_{1,2}^1, z_{o2}, z_{s2}, z_{2,3}^1) \quad (5.22)$$

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Function F has similar format as those in (5.10) and (5.17), except that the highest order in numerator is 8 and that in the denominator is 5. By equalizing (5.21) and (5.8) with $n = 3$ and $q = 5$, five equations can be derived in terms of six variables. In this aspect, the element values are determined in such a way that the frequency response derived from the network is approaching the targeted one. It needs to be noticed that the transmission zeros appeared in multiple integer times of 180° are 5th-order transmission zeros. Thus, if both single stage and three-stage bandpass filters are designed under the same specifications, *i.e.* ripple constant, lower cutoff phase, and same number of transmission poles, the rejection skirts of the three-stage filter definitely becomes sharper at the passband edges.

One design example is given to illustrate the three-stage case with $\theta_c = 46^\circ$ at 3 GHz. The relevant transmission-line model shown in Fig. 5.12 is implemented using hybrid microstrip line and slotline structures as the single stage cases. The schematic layout is depicted in Fig. 5.13 with all dimensions indicated, and design specifications are given in Table 5.6. The series open stubs are all constructed by microstrip lines with open-circuited ends, while the shunt short stubs are all formed by slotlines in the ground. The uniform lines between two sets of stubs can be either microstrip line or slotline. From the multiple-mode resonator point of view [57], the band-pass filter of this constituted filter is realized by three such resonators, *i.e.*, one microstrip line resonator and two slotline resonators, which are coupled via microstrip-to-slotline transitions. The first two resonant modes of these three resonators contribute to six transmission poles. Plus two poles from input/output coupling structures, this filter can achieve eight transmission poles in the passband in total. Due to the physical constraint in achieving specified values of

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characteristic impedances within the filter layout, the terminal impedances at the two references are chosen as $70\ \Omega$. To connect this filter with two $50\text{-}\Omega$ coaxial cables in measurement, two quarter-wavelength transformers are additionally designed at the central frequency and they are installed in the two main feed lines as shown in Fig.5.13.

To verify the theoretical predictions, the designed three-stage band-pass filter is finally fabricated and measured. Fig. 5.14(a) and (b) shows the top- and bottom-view photographs of the fabricated circuits. The measured frequency responses and group delays are plotted in Fig. 5.15(a) and (b), respectively. Basically, the measured results agree with the simulations well over the frequency range of 1.0 to 15.0 GHz, achieving a wide passband with the fractional bandwidth of 101.6% at 5.51 GHz. The insertion loss is lower than 2.32 dB and the return loss is larger than 7.2 dB in the core passband. The measured group delays are basically flat within the passband while they tremendously vary as the frequency approaches the passband edges as exhibited in the reported bandpass filters. The peak appears around 10.5 GHz in the measurement result is probably introduced by the slight fabrication differences among all the shunt or series stubs in electrical length.

In general, the proposed transmission-line model is more attractive in the design of bandpass filters with ultra-wide bandwidth due to its inherent high-pass property. By using the hybrid microstrip line and slotline to implement the presented filter networks,

TABLE 5.6
CIRCUIT PARAMETERS FOR EIGHT-POLE FILTER ($\theta_c = 46^\circ$ at 3 GHz)

Stub line impedance	Connecting line impedance
$Z_{o1} = 86.80\ \Omega$	$Z_{1,2}^1 = 75.46\ \Omega$
$Z_{s1} = 74.20\ \Omega$	$Z_{2,3}^1 = 70\ \Omega$
$Z_{o2} = 94.43\ \Omega$	$Z_0 = 70\ \Omega$
$Z_{s2} = 50.40\ \Omega$	

Fig. 5.14. Photographs of fabricated eight-pole ultra-wideband bandpass filter in Fig. 4.12. (a) Top-view. (b) Bottom-view.

the range of lower cutoff phase θ_c is restricted to 20° - 50° due to the limitation of realizable characteristic impedances of the transmission lines. Therefore, the range of applicable fractional bandwidth for the proposed filter structure can be estimated as about 85% to 155% under the specified ripple level and pole number in the desired passband. In addition, characteristic impedance and effective permittivity of the slotline involved in this proposed filter are functions of frequency within the passband. It is one primary factor that causes frequency shift in a high frequency range between the synthesis-predicted results and EM-simulated or measured results. To minimize this parasitic effect, low-impedance slotline with narrow slot width is preferred in implementation if possible. Under the restriction in synthesis procedure that requires the characteristic impedance and effective permittivity to be constant, these two parameters for the slotline at center frequency of the passband should be selected.

On the other hand, as mentioned earlier, there exist some visible derivations between the synthesis-derived results and EM-simulated or measured results. They are primarily attributed by the common drawback of all the synthesis design procedures in microwave

Fig. 5.15. Eight-pole ultra-wideband bandpass filter in Fig. 4.12. (a) Predicted and measured S-parameters. (b) Predicted and measured group delays.

planar filters, i.e., no ability in taking into account all of the unexpected dynamic effects inclusive of frequency dispersion, discontinuity effects, conductor/material/radiation losses and so on. However, these effects have been appearing in both EM-simulated and experimental results. As shown in Figs. 5.6, 5.8, 5.11 and 5.15, these two sets of results are well matched with each other. Even though it is theoretically approximate, the synthesis procedure is always highly desired in filter design. It is because that the synthesis approach not only gives people a deeply physical insight into the operating principle for the developed microwave circuits but also remarkably reduces the

computation time and design circles especially for the high-order filter structures with many unknown dimensions. In this case, all the dimensions obtained from the synthesis approach can be used as the initial dimensions for more accurate EM-based simulation, tuning and optimization.

5.3 Conclusion

In this Chapter, a direct synthesis approach for a class of ultra-wideband single stage and multi-stage bandpass filters is systematically presented. The transfer functions derived from the exact networks are firstly formulated to realize the Chebyshev equal-ripple responses in the wide passband and to explicitly determine the normalized values of transmission-line elements under the required specifications. Three single-stage and one three-stage filter prototypes are extensively investigated and they are then designed and implemented on hybrid slotline and microstrip line structures. Finally, four filter circuits are fabricated and measured. As evidently confirmed in simulation and experiment, the developed synthesis method provides us with an efficient, effective and accurate solution in design and exploration of a class of ultra-wideband band-pass filters with composite short/open stubs.

CHAPTER 6

Conclusions and Recommendations

6.1 Conclusions

In this dissertation, a dual network of the conventional high-pass prototype with series open/short-circuited stubs is investigated and synthesized to construct low/high-pass filters. The insertion loss function is derived directly from the proposed transmission-line model and regulated to exhibit Chebyshev equal-ripple responses in the passband. By equalizing the derived insertion loss function with the ideal theoretical generating functions, the element values involved in the transmission-line model are able to be calculated and tabulated. The presented networks are realized and implemented on the hybrid coplanar-waveguide and slotline structures or composite microstrip line and

slotline structures. Then the designed filters are optimized and fabricated to experimentally confirm and validate the theoretical predictions. To suppress the higher order harmonic bands and extend the upper stopband, the uniform transmission lines between the adjacent stubs can be replaced by the periodic structures with slow-wave behavior. This network provides the designers an alternative choice of constructing UWB bandpass filters.

With the non-uniform MMR as a basic structure, the stub-/slot-loaded MMR is proposed to construct UWB bandpass filter with its first few resonant modes. The effects of loading stubs/slots at specified positions of an open-circuited microstrip line MMR are studied in detail. This property is then utilized to locate the resonances within the designed passband. With the external coupling structures, the bandpass filters with ultra-wide bandwidth can be formed. The two-stage slot-loaded MMR UWB bandpass filter is also investigated with the help of mode graphs. The filter selectivity is enhanced by cascading more resonators and the upper stopband is effectively widened. Another way to sharpen the rejection skirts is to generate two transmission zeros at both the upper and lower passband edges. It is done by loading two open-circuited non-uniform stubs at the feed lines. The locations of two zeros are controlled by the impedance ratio of the two sections. The designed filters are fabricated and measured to experimentally exhibit the good UWB filtering responses.

The MMR theory can also be applied on other transmission-line structures. The UWB bandpass filter with slotline resonator and microstrip-to-slotline transitions is proposed and analyzed. With the first few resonances contributed by the uniform/non-uniform slotline resonator, plus the microstrip-to-slotline transitions at two ports as

distributive coupling structure, the bandpass filter with controllable ultra-wide bandwidth is designed. To further improve the filter performance, cross-coupling path is introduced by twisting the slotline resonator to a W or bowl shape. Two transmission zeros are created by such geometrical variation and they enhance the filter selectivity greatly. Low-pass structures implemented on the ground plane below the feed lines or on the microstrip feed lines are installed at two ports. The upper stopband performance is effectively improved without affecting the in-band frequency responses. The prototype filters are also fabricated and measured to validate the simulated results.

The general transmission-line model with hybrid series and shunt stubs is proposed and analyzed with the synthesis approach. The insertion loss function is derived from the presented network and regulated to exhibit Chebyshev equal-ripple characteristics in the passband. The impedances of the stubs and connecting lines can be calculated by equalizing the derived insertion loss function with the ideal target function and the filter dimensions can be determined consequently without tedious EM-based simulations. By increasing the number of transmission-line sections between two sets of hybrid stubs, one can increase the number of transmission poles in the main passband. The proposed network is implemented on hybrid microstrip line and slotline structures and can be used to explain the slotline resonator UWB filter from synthesis design point of view. With the replacement of uniform connecting lines with slow-wave lines that have electrical properties similar to those of the original transmission line, the harmonic bands can be effectively suppressed and the upper stopband is widened. The transmission-line model of the multiple-stage bandpass filter is also presented and studied with similar synthesis procedure. The advantage of the single and multiple-stage bandpass filters over the

traditional high-pass prototype is that they provide higher order transmission zeros and improve the filter selectivity consequently. More importantly, the proposed prototypes give the designers alternative choice to design UWB bandpass filters with tabulated design parameters as reference.

Since several UWB bandpass filters have been proposed and designed in this dissertation, there is a question on how to select these filters for an actual application. Generally speaking, if a UWB bandpass filter with the specified in-band bandwidth and roll-off skirt near the cut-off frequency edges, the high-pass prototype with series stubs or with hybrid stubs are always considered in the first priority. It is because the initial dimensions of these prototype filters can be straightforwardly decided via the presented synthesis approach. For a real filter that needs to have a wide upper stopband and/or smaller size, these two prototypes can be served as the basic structures with more added feature. The uniform connecting line between adjacent stubs can be replaced by high-stop filtering structures, such as electromagnetic bandgap (EBG), to suppress unexpected high harmonic bands, thereby extending the upper stopband. The total circuitry area can be simply reduced by geometrical rearrangement. Of course, the proposed slot-loaded MMR UWB filter can also get size reduction due to its slow-wave property. Next, based on the initial dimensions derived above, various bandpass filters with the specified fractional bandwidth need to be numerically optimized with virtue to full-wave simulation on its overall layout as carried out in many literatures. The shaper rejection skirts can be mainly achieved through cascading more MMRs. As the cost, this direct approach unfortunately increases the insertion loss and overall circuit size. If an UWB bandpass filter with high selectivity is required, multi-stage hybrid stub based UWB bandpass filter and stub-

loaded T-slot MMR based UWB bandpass filter are good candidates. If rejection or roll-off rate is not a critical issue while compact size is desirable in priority, the stub-loaded MMR based filter can be used. To conclude, there are always pros and cons for the filter design and the final selection much depends on which feature has the highest priority.

6.2 Future Recommendations

All filter prototypes mentioned so far are composed of transmission-line structures with uniform characteristic impedances, such as stubs and connecting lines. The performance of these networks with non-uniform structures becomes the topic under exploration. With the replacement of uniform shunt open-ended stubs [128], the conventional quasi-low-pass filter was used to design band-stop filters with wider stopband bandwidth. The introduction of non-uniform stubs or connecting lines will give additional design freedom to investigate more desired features. With the modern synthesis analysis procedure, the transfer function of the proposed transmission-line network with non-uniform structures could be established. To regulate this transfer functions to exhibit the desired responses, a targeted transfer function is needed. However, with this complex structure, there might not be a well-developed function. Modifications on the existing functions might be required. Once the filter prototype is theoretically developed, it will be implemented on different transmission-line structures. Multi-layer structures, such as liquid crystal polymer and low temperature co-fired ceramic might be used. With geometrical variations and making use of the vertical axes to achieve filter compactness, additional features as well as integration with other circuit components in the system. The synthesis design method for UWB band-pass filter can also be extended

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to explore other types of passive devices for UWB applications, such as balanced to unbalanced transformers (Balun) and coupler. The development of these devices with wide operating frequency range is rather limited [100] [128]. The analytical procedure will be similar but much more complicated for the devices with multiple ports.

Appendix I

Element Values for Equal-Ripple Low-Pass Filters

0.1 dB ripple				
<i>n</i>	θ_c	$a_1 = a_2$	b_1	
2	5°	11.69756	0.15146	
	10°	5.66789	0.29908	
	15°	3.58615	0.43928	
	20°	2.50026	0.56900	
	25°	1.82159	0.68590	
	30°	1.35462	0.78848	
	35°	1.01548	0.87609	
	40°	0.76165	0.94893	
	45°	0.56867	1.00784	
	50°	0.42090	1.05421	
	55°	0.30755	1.08966	
	60°	0.22070	1.11596	
	65°	0.15436	1.13481	
<i>n</i>	θ_c	$a_1 = a_3$	a_2	$b_1 = b_2$
3	5°	13.03891	22.41496	0.126973
	10°	6.36736	10.8864	0.25209
	15°	4.07985	6.91021	0.37357
	20°	2.89266	4.83239	0.48971
	25°	2.15017	3.52498	0.59901
	30°	1.63468	2.61402	0.70013
	35°	1.25338	1.94064	0.79197
	40°	0.96020	1.42628	0.87370
	45°	0.72969	1.02772	0.94472
	50°	0.54659	0.71896	1.00474
	55°	0.40104	0.48294	1.05378
	60°	0.28620	0.30726	1.09219
	65°	0.19690	0.18175	1.12075

Appendix II

Element Values for Equal-Ripple Low-Pass Filters

1.0 dB ripple				
<i>n</i>	θ_c	$a_1 = a_2$		b_1
2	5°	23.02793		0.17513
	10°	11.27560		0.34762
	15°	7.25820		0.51493
	20°	5.18102		0.67470
	25°	3.88683		0.82483
	30°	2.99102		0.96355
	35°	2.32941		1.08947
	40°	1.82026		1.20162
	45°	1.41814		1.29948
	50°	1.09562		1.38295
	55°	0.83488		1.45234
	60°	0.62356		1.50834
	65°	0.45263		1.55192
<i>n</i>	θ_c	$a_1 = a_3$	a_2	$b_1 = b_2$
3	5°	24.31704	34.11823	0.15967
	10°	11.9399	16.6584	0.31760
	15°	7.72082	10.6690	0.47211
	20°	5.54545	7.55688	0.62154
	25°	4.19212	5.60625	0.76435
	30°	3.25452	4.24734	0.89914
	35°	2.55919	3.23762	1.02462
	40°	2.01991	2.45711	1.13967
	45°	1.58910	1.84006	1.24336
	50°	1.23848	1.34762	1.33494
	55°	0.95016	0.95536	1.41388
	60°	0.71229	0.64698	1.47987
	65°	0.51665	0.41075	1.53289

Appendix III

Element Values for Equal-Ripple Four-Pole Band-Pass Filters

0.1 dB ripple			
θ_c	z_a	z_b	z_c
20	0.38585	4.05379	1.16418
25	0.49734	3.03311	1.16418
30	0.61914	2.35026	1.16418
35	0.75475	1.85853	1.16418
40	0.90877	1.48609	1.16418
45	1.08771	1.19408	1.16418
50	1.30138	0.95874	1.16418
55	1.56500	0.76480	1.16418
60	1.90387	0.60215	1.16418
65	2.36342	0.46381	1.16418

Appendix IV

Element Values for Equal-Ripple Five-Pole Band-Pass Filters

0.1 dB ripple			
θ_c	z_a	z_b	z_c
20	0.39449	3.49064	1.20995
25	0.51008	2.53918	1.24030
30	0.63669	1.91833	1.28164
35	0.77772	1.48569	1.33669
40	0.93779	1.16933	1.40952
45	1.12368	0.92963	1.50567
50	1.34536	0.74259	1.63383
55	1.61858	0.59291	1.80736
60	1.96948	0.47026	2.04865
65	2.44506	0.36737	2.39813

Bibliography

- [1] A. A. Oliner, "Historical perspectives on microwave field theory," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, no. 9, pp. 1022–1045, Sep. 1984.
- [2] H. Sobol and K. Tomiyasu, "Milestones of microwaves," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 3, pp. 594–611, Mar. 2002.
- [3] K. C. Gupta, Tatsuo Itoh and A. A. Oliner, "Microwave and RF education—past, present, and future," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 3, pp. 1006–1014, Mar. 2002.
- [4] P. Staecker, "Microwave industry outlook—overview," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 3, pp. 1034–1036, Mar. 2002.
- [5] D. M. Pozar, *Microwave Engineering*, 2nd edition, John Wiley & Sons, 1998.
- [6] R. E. Collin, *Foundations for Microwave Engineering*, 2nd Edition, McGraw Hill, 1992.
- [7] J. H. Howe, "Microwave integrated circuits – an historical perspective," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, no. 9, pp. 991–996, Sep. 1984.
- [8] J. D. N. Mcquiddy, J. W. Wassel, J. B. Lagrange and W. Wisseman, "Monolithic microwave integrated circuits—an historical perspective," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, no. 9, pp. 997–1008, Sep. 1984.
- [9] E. C. Niehenke, R. A. Pucel and I. J. Bahl, "Microwave and millimeter-wave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 3, pp. 846–857, Mar. 2002.

-
- [10] R. M. Barrett, "Microwave printed circuits—A historical survey," *IRE Trans. Microwave Theory Tech.*, vol. MTT-3, pp. 1–9, Mar. 1955.
  - [11] T. Itoh, *Planar Transmission Line Structures*, New York: IEEE Press, 1987.
  - [12] R. M. Barrett, "Microwave printed circuits—The early years," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, pp. 983–990, Sept. 1984.
  - [13] D. D. Grieg and H. F. Engelmann, "Microstrip – A new transmission technique for the kilomegacycle range," *Proc. IRE*, vol. 40, pp. 1644–1650, Dec. 1952.
  - [14] F. Assadourian and E. Rimai, "Simplified theory of microstrip transmission systems," *Proc. IRE*, vol. 40, pp. 1651–1657, Dec. 1952.
  - [15] S. B. Cohn, "Slotline on a dielectric substrate," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-17, pp. 768–778, Oct. 1969.
  - [16] K. C. Gupta, R. Garg, I. Bahl and P. Bhartia, *Microstrip Lines and Slotlines*, 2nd edition, Norwood, MA: Artech House, 1996.
  - [17] H. A. Wheeler, "Transmission line properties of a stripline between parallel planes," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-26, pp. 866–876, Nov. 1978.
  - [18] R. N. Simons, *Coplanar Waveguide Circuits, Components, and Systems*, John Wiley & Sons, 2001.
  - [19] G. L. Matthaei, L. Young, and E. M. T. Jones, *Microwave Filters, Impedance-Matching Networks and Coupling Structures*. New York: McGraw-Hill, 1964.
  - [20] E. G. Cristal and S. Frankel, "Hairpin-line and hybrid hairpin-line/halfwave parallel-coupled-line filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 719–728, Nov. 1972.

-
- [21] M. Sagawa, K. Takahashi, and M. Makimoto, "Miniaturized hairpin resonator filters and their application to receiver front-end MIC's," *IEEE Trans. Microwave Theory Tech.*, vol. 37, pp. 1991–1997, Dec. 1989.
  - [22] J. S. Hong and M. J. Lancaster, "Cross-Coupled Microstrip Hairpin-Resonator Filters," *IEEE Trans. Microwave Theory Tech.*, vol. 46, no. 1, pp. 118–122, Jan. 1998.
  - [23] J.-T. Kuo and H.-S. Cheng, "Design of quasi-elliptic function filters with a dual-passband response," *IEEE Microw. Wireless Compon. Lett.*, vol. 14, no. 10, pp. 472–474, Oct. 2004.
  - [24] G. L. Matthaei, "Interdigital band-pass filter," *IEEE Trans. Microwave Theory Tech.*, vol. 10, no. 6, pp. 479 – 491, Nov. 1962.
  - [25] S.W. Ting, K.W. Tam, and R.P. Martins, Novel interdigital microstrip bandpass filter with improved spurious response, *IEEE ISCAS 1* (2004), 984–987.
  - [26] J. R. Crute and L. E. Davis, "A compact microstrip interdigital stepped impedance bandpass filter with enhanced stop band," *Microw. Opt. Technol. Lett.*, vol. 34, no. 5, pp. 336–340, Sep. 2002.
  - [27] I. Wolff, "Microstrip bandpass filter using degenerate modes of a microstrip ring resonator," *Electron. Lett.*, vol. 8, no. 12, pp. 302–303, Jun. 1972.
  - [28] U. Karacaoglu, D. Sanchez-Hernandez, I. D. Robertson, and M. Guglielmi, "Harmonic suppression in microstrip dual-mode ring-resonator bandpass filters," in *IEEE MTT-S Int. Microw. Symp. Dig.*, 1996, pp. 1635–1638.
  - [29] A. Görür, "A novel dual-mode bandpass filter with wide stopband using the properties of microstrip open-loop resonator," *IEEE Microw. Wireless Compon. Lett.*, vol. 12, no. 10, pp. 386–388, Oct. 2002.

-
- [30] S. W. Fok, P. Cheong, K. W. Tam, and R. P. Martins, "Microstrip dual-mode bandpass filter design with simultaneous size reduction and spurious response suppression," in *IEEE MTT-S Int. Microw. Symp. Dig.*, 2005, pp. 2175–2178.
  - [31] J.-T. Kuo and C.-Y. Tsai, "Periodic stepped-impedance ring resonator (PSIRR) bandpass filter with a miniaturized area and desirable upper stopband characteristics," *IEEE Trans. Microw. Theory Tech.*, vol. 54, no. 3, pp. 1107–1112, Mar. 2006.
  - [32] L. Zhu, P. Wecowski, and K. Wu, "New planar dual-mode filter using cross-slotted patch resonator for simultaneous size and loss reduction," *IEEE Trans. Microw. Theory Tech.*, vol. 47, no. 5, pp. 650–654, May 1999.
  - [33] A. Cassinese, F. Palomba, G. Pica, and A. Andreone, "Dual mode crossslotted filters realized with superconducting films," *Appl. Phys. Lett.*, vol. 77, no. 26, pp. 4407–4409, Dec. 2000.
  - [34] L. Zhu, B. C. Tan, and S. J. Quek, "Miniaturized dual-mode bandpass filter using inductively loaded cross-slotted patch resonator," *IEEE Microw. Wireless Compon. Lett.*, vol. 15, no. 1, pp. 22–24, Jan. 2005.
  - [35] W.-H. Tu and K. Chang, "Miniaturized dual-mode bandpass filter with harmonic control," *IEEE Microw. Wireless Compon. Lett.*, vol. 15, no. 12, pp. 838–840, Dec. 2005.
  - [36] R. Fontana, "A Brief History of UWB Communications," online article, <http://www.multispectral.com/history.html>
  - [37] N. Faranak, *Ultra-Wideband Communications: Fundamentals and Applications*. Prentice Hall, 2006.

-
- [38] G. R. Aiello and G. D. Rogerson, "Ultra-wideband wireless systems," *IEEE Microwave Mag.*, vol. 4, no. 2, pp. 36–47, Jun. 2003.
- [39] L. Q. Yang and G. B. Giannakis, "Ultra-wideband communications: an idea whose time has come," *IEEE Signal Processing Mag.*, vol. 21, no. 11, pp. 26–54, Nov. 2004.
- [40] D. Porcino and W. Hirt, "Ultra-wideband radio technology: Potential and challenges ahead," *IEEE Commun. Mag.*, vol. 41, pp. 66–74, Jul. 2003.
- [41] S. Roy, J.R. Foerster, V.S. Somayazulu, and D.G. Leeper, "Ultra-wideband radio design: The promise of high-speed, short-range wireless connectivity," *Proc. IEEE*, vol. 92, no. 2, pp. 295–311, Feb. 2004.
- [42] R. J. Fontana, "Recent system applications of short-pulse ultra-wideband (UWB) technology," *IEEE Trans. Microwave Theory Tech.*, vol. 52, no. 9, pp. 2087–2104, Sep. 2004.
- [43] "Revision of part 15 of the commission's rules regarding ultra-wideband transmission systems," in *First note and Order, Federal Communications Commission*, ET-Docket 98-153, Feb. 14, 2002.
- [44] R. Levy and L. F. Lind, "Synthesis of symmetrical branch-guide directional couplers," *IEEE Trans. Microw. Theory and Tech.*, vol. MTT-16, pp.80-89, Feb. 1968.
- [45] W. -T. Wong, Y. -S. Lin, C. -H. Wang, and C. H. Chen, "Highly selective microstrip bandpass filters for ultra-wideband (UWB) applications," in *Proc. Asia-Pacific Microw. Conf.*, vol. 5, Dec. 2005, pp. 2850 -2853.

-
- [46] R. Gomez-Garcia and J. I. Alonso, "Systematic method for the exact synthesis of ultra-wideband filtering responses using high-pass and low-pass sections," *IEEE Trans Microw. Theory Tech.*, vol. 54, no. 10, pp. 3751–3764, Oct. 2006.
- [47] J. Garcia-Garcia, J. Bonache, and F. Martin, "Application of electromagnetic bandgaps to the design of ultra-wide bandpass filters with good out-of-band performance," *IEEE Trans Microw. Theory Tech.*, vol. 54, no. 12, pp. 4136–4140, Dec. 2006.
- [48] H. Shaman and J. -S. Hong, "A novel ultra-wideband (UWB) bandpass filter (BPF) with pairs of transmission zeroes," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 2, pp. 121–123, Feb. 2007.
- [49] H. Shaman and J. -S. Hong, "Input and output cross-coupled wideband bandpass filter," *IEEE Trans Microw. Theory Tech.*, vol. 55, no. 12, pp. 2562–2568, Dec. 2007.
- [50] M. Uhm, K. Kim, D. S. Filipovic, "Ultra-wideband bandpass filters using quarter-wave short-circuited shunt stubs and quarter-wave series transformers" *IEEE Microw. Wireless Compon. Lett.*, vol. 18, no. 10, pp. 668–670, Oct. 2008.
- [51] Z.-C. Hao and J.-S. Hong, "Ultra-wideband bandpass filter using multilayer liquid-crystal-polymer technology," *IEEE Trans. Microw. Theory Tech.*, vol. 56, no. 9, pp. 2095–2100, Aug. 2008.
- [52] Z.-C. Hao and J.-S. Hong, "Ultra wideband bandpass filter using embedded stepped impedance resonators on multilayer liquid crystal polymer substrate," *IEEE Microw. Wireless Compon. Lett.*, vol. 18, no. 9, pp. 581–583, Sep. 2008.

-
- [53] Z.-C. Hao and J.-S. Hong, "Compact ultra-wideband bandpass filter using broadside coupled hairpin structures on multilayer liquid crystal polymer substrate," *Electron. Lett.*, vol. 44, no. 20, pp. 1197–1198, Sep. 2008.
- [54] W. Menzel, L. Zhu, K. Wu, and F. Bogelsack, "On the design of novel compact broad-band planar filters," *IEEE Trans. Microw. Theory Tech.*, vol. 51, no. 2, 364–370, Feb. 2003.
- [55] L. Zhu, H. Bu, and K. Wu, "Broadband and compact multi-mode microstrip bandpass filters using ground plane aperture technique," *IEE Proc. –Microw. Antennas Propag.*, vol. 149, no. 1, 71–77, Feb. 2002.
- [56] L. Zhu, H. Bu, K. Wu and M. S. Leong, "Miniaturized multi-pole broadband microstrip bandpass filter: concept and verification," in *30th European Microwave Conf. Proc.*, vol.3, Paris, pp.334–337, Oct. 2000.
- [57] L. Zhu, S. Sun, and W. Menzel, "Ultra-wideband (UWB) bandpass filters using multiple-mode resonator," *IEEE Microw. Wireless Compon. Lett.*, vol. 15, no. 11, pp. 796–798, Nov. 2005.
- [58] H. Wang, L. Zhu, and W. Menzel, "Ultra-wideband bandpass filter with hybrid microstrip/CPW structure," *IEEE Microw. Wireless Compon. Lett.*, vol. 15, no. 12, pp. 844–846, Dec. 2005.
- [59] J. Gao, L. Zhu, W. Menzel, and F. Bogelsack, "Short-circuited CPW multiple-mode resonator for ultra-wideband (UWB) bandpass filter," *IEEE Microw. Wireless Compon. Lett.*, vol. 16, no. 3, pp. 104–106, Mar. 2006.

-
- [60] J. Gao and L. Zhu, "Asymmetric parallel-coupled CPW stages for harmonic suppressed 1/4 bandpass filters," *Electron. Lett.*, vol. 40, no. 18, pp. 1122–1123, Sept. 2004.
  - [61] R. Li and L. Zhu, "Compact UWB bandpass filter using stub-loaded multiple-mode resonator," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, No.1, January, 2007, pp.40-42.
  - [62] S. Sun and L. Zhu, "Capacitive-ended interdigital coupled lines for UWB bandpass filters with improved out-of-band performance," *IEEE Microw. Wireless Compon. Lett.*, vol. 16, no. 8, pp. 440–442, Aug. 2006.
  - [63] N. Thomson and J. -S. Hong, "Compact ultra-wideband microstrip/coplanar waveguide bandpass filter," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 3, pp. 184-186, Mar. 2007.
  - [64] K. Li, D. Kurita, and T. Matsui, "An ultra-wideband bandpass filter using broadside-coupled microstrip-coplanar waveguide structure," in *IEEE MTT-S Int. Dig.*, Jun. 2005, pp. 675-678.
  - [65] K. Li and J. -S Hong, "Modeling of an ultra-wideband bandpass filtering structure," *Proceeding of APMC2006*, vol. 1, pp. 37-40, 2006.
  - [66] K. Li, Y. Yamamoto, T. Matsui and O. Hashimoto, "An ultra-wideband(UWB) bandpass filter using broadside-coupled structure and shunt stub with chip capacitor," *Proceeding of APMC2006*, vol. 1, pp. 41-44, 2006.
  - [67] R. Li, and L. Zhu, "Ultra-wideband (UWB) bandpass filters with hybrid microstrip/slotline structures," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 11, pp. 778–780, Nov. 2007.

-
- [68] R. Li and L. Zhu, "Ultra-wideband microstrip-slotline bandpass filter with enhanced rejection skirts and widened upper stopband," *Electron. Lett.*, vol. 43, no. 24, pp. 1368–1369, Nov. 2007.
- [69] S. W. Wong and L. Zhu, "EBG-embedded multiple-mode resonator for UWB bandpass filter with improved upper-stopband performance," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 6, pp. 421–423, Jun. 2007.
- [70] S. W. Wong and L. Zhu, "Ultra-wideband bandpass filters with improved out-of-band behavior via embedded electromagnetic-bandgap multimode resonators," Accepted by *IET Microwaves, Antennas & Propagation*, vol. 2, no. 8, pp. 854–862, Sep. 2008.
- [71] J.-W. Baik, S.-M. Han, C. Jeong, J. Jeong and Y. -S. Kim, "Compact ultra-wideband bandpass filter with EBG structure," *IEEE Microw. Wireless Compon. Lett.*, vol. 18, no. 10, pp. 671–673, Oct. 2008.
- [72] T. B. Lim, S. Sun, and L. Zhu, "Compact ultra-wideband bandpass filter using harmonic-suppressed multiple-mode resonator," *Electron. Lett.*, vol. 43, no. 22, pp. 1205–1206, Oct. 2007.
- [73] C. -W. Tang and M. -G. Chen, "A microstrip ultra-wideband bandpass filter with cascaded broadband bandpass and bandstop filters," *IEEE Trans. Microw. Theory Tech.*, vol. 55, no. 11, 2412–2418, Nov. 2007.
- [74] A. Balalem, W. Menzel, J. Machac, and A. Omar, "A simple ultra-wideband suspended stripline bandpass filter with very wide stop-band," *IEEE Microw. Wireless Compon. Lett.*, vol. 18, no. 3, pp. 170–172, Mar. 2008.

-
- [75] H. Shaman and J. -S. Hong, "Ultra-Wideband (UWB) bandpass filter with embedded band notch structures," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 3, pp. 193-195, Mar. 2007.
- [76] H. Shaman and J. -S. Hong, "Asymmetric parallel-coupled lines for notch implementation in UWB filters," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 7, pp. 516-518, Jul. 2007.
- [77] K. Li, D. Kurita, T. Matsui, "Dual-band ultra-wideband bandpass filter," in *IEEE MTT-S Int. Dig.*, Jun. 2007, pp. 1193-1196.
- [78] S. W. Wong and L. Zhu, "Implementation of compact UWB bandpass filter with a notch-band," *IEEE Microw. Wireless Compon. Lett.*, vol. 18, no. 1, pp. 10-12, Jan. 2008.
- [79] W. Menzel and P. Feil, "Ultra-wideband (UWB) filter with WLAN notch," in *36th European Microw. Conf.*, Sep. 2006, 595-598.
- [80] G. -M. Yang, R. Jin, C. Vittoria, V. G. Harris, and N. X. Sun, "Small ultra-wideband (UWB) bandpass filter with notched band," *IEEE Microw. Wireless Compon. Lett.*, vol. 18, no. 3, pp. 176-178, Mar. 2008.
- [81] K. Li, D. Kurita, T. Matsui, "UWB bandpass filters with multi notched bands," in *36th European Microw. Conf.*, Sept. 2006, 591-594.
- [82] H. Ishida and K. Araki, "Design and analysis of UWB bandpass filter with ring filter," in *IEEE MTT-S Int. Dig.*, vol. 3, Jun. 2004, pp. 1307-1310.
- [83] C. -L. Hsu, F. -C. Hsu, and J. -T. Kuo, "Microstrip bandpass filters for ultra-wideband (UWB) wireless communications," in *IEEE MTT-S Int. Dig.*, Jun. 2005, pp. 679-682.

-
- [84] W. -T. Wong, Y. -S. Lin, C. -H. Wang and C. H. Chen, "Highly selective microstrip bandpass filters for ultra-wideband (UWB) applications," in *Proc. Asia-Pacific Microw. Conf.*, vol. 5, Dec. 2005, pp. 2850 -2853.
- [85] W. Menzel, M. S. R. Tito, and L. Zhu, "Low-loss ultra-wideband (UWB) filters using suspended stripline," in *Proc. 2005 Asia-Pacific Microw. Conf.*, vol. 4, Dec. 2005, pp. 2148-2151.
- [86] J. Lee, Y. -J. Park, M. Kim, C. Yoon, J. Kim, and K. -H. Kim, "System-on-package ultra-wideband transmitter using CMOS impulse generator," *IEEE Trans. Microw. Theory Tech.*, vol. 54, no. 4, pp. 1667–1774, Apr. 2006.
- [87] M. K. Mandal and S. Sanyal, "Compact wideband bandpass filter," *IEEE Microw. Wireless Compon. Lett.*, vol. 16, no. 1, pp. 46–48, Jan. 2006.
- [88] K. M. Shum, W. T. Luk, C. H. Chan and Q. Xue, "A UWB bandpass filter with two transmission zeros using a single stub with CMRC," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 1, pp. 43-45, Jan. 2007.
- [89] M. Mokhtaari, J. Bornemann and S. Amari, "Folded compact ultra-wideband stepped-impedance resonator filters," in *IEEE MTT-S Int. Dig.*, Jun. 2007, pp. 747–750.
- [90] Y.-C. Chiou, J.-T. Kuo and E. Cheng, "Broadband quasi-Chebyshev bandpass filter with multimode stepped-impedance resonators (SIRs)," *IEEE Trans. Microw. Theory Tech.*, vol. 54, pp. 3352-3358, Aug. 2006.
- [91] J.-T. Kuo, Y.-C. Chiou and E. Cheng, "High selectivity ultra-wideband (UWB) multimode stepped-impedance resonators (SIRs) bandpass filter with two-layer

- broadside-coupled structure,” in *Proc. Asia-Pacific Microw. Conf.*, Dec. 2007, pp.1–4.
- [92] A. M. Abbosh, “Planar bandpass filters for ultra-wideband applications,” *IEEE Trans Microw. Theory Tech.*, vol. 55, no. 10, pp. 2262–2269, Oct. 2007.
- [93] T.-N. Kuo, S.-C. Lin, and C. H. Chen, “Compact ultra-wideband bandpass filters using composite microstrip-coplanar-waveguide structure,” *IEEE Trans. Microw. Theory Tech.*, vol. 54, no. 10, pp. 3772–3778, Oct. 2006.
- [94] J. -S. Hong and K. Li, “Recent development of ultra-wideband (UWB) filters,” *IEEE 2007 International Symposium on Microwave, Antenna, Propagation and EMC Technologies for Wireless Communications, MAPE*, art, no. 4393646, pp. 442-445.
- [95] S. Sun and L. Zhu, “Multiple-mode-resonator-based bandpass filters for ultra-wideband transmission systems,” Accepted by *IEEE Microwave Magazine*, in October 2008.
- [96] K. Kiminami, A. Hirata, and T. Shiozawa, “Double-sided printed bow-tie antenna for UWB communications,” *IEEE Antennas and Wireless Propagation Lett.*, vol. 3, no. 1, pp. 152–153, 2004.
- [97] J. Liang, C. C. Chiau, X. Chen and C. G. Prini, “Study of a printed circular disc monopole antenna for UWB systems,” *IEEE Trans. Antennas Propag.*, vol. 53, no. 11, pp. 3500–3504, Nov. 2005.
- [98] A. Bevilacqua and A. M. Niknejad, “An ultra-wideband CMOS low-noise amplifier for 3.1-10.6-GHz wireless receivers,” *IEEE J. Solid-State Circuits*, vol. 39, no. 12, pp. 2259–2268, Dec. 2004.

-
- [99] A. Ismail and A. A. Abidi, "A 3-10-GHz low-noise amplifier with wideband LC-ladder matching network," *IEEE J. Solid-State Circuits*, vol. 39, no. 12, pp. 2269–2277, Dec. 2004.
- [100] A. M. Abbosh and M. E. Bialkowski, "Design of compact directional couplers for UWB applications," *IEEE Trans. Microw. Theory Tech.*, vol. 55, no. 2, pp. 189–194, Feb. 2007.
- [101] A. M. Abbosh, "A compact UWB three-way power divider," *IEEE Microw. Wireless Compon. Lett.*, vol. 17, no. 8, pp. 598–560, Aug. 2007.
- [102] P. I. Richard, "Resistor-transmission line circuits," *Proc. of the IRE*, vol. 36, pp. 217–220, Feb. 1948.
- [103] H. Ozaki, and J. Ishii, "Synthesis of a class of strip-line filters," *IRE Trans. Circuit Theory*, vol. 5, no. 2, pp. 104–109, Jun. 1958.
- [104] H. Ozaki, and J. Ishii, "Synthesis of transmission-line networks and the design of UHF filters," *IRE Trans. Circuit Theory*, vol. 2, no. 4, pp. 325–336, Dec. 1955.
- [105] H. J. Riblet, "The application of a new class of equal-ripple functions to some familiar transmission-line problems," *IEEE Trans. Microw. Theory and Tech.*, vol. 12, no. 4, pp. 415–421, Jun. 1964.
- [106] C. B. Sharpe, "A general tchebycheff rational function," *Proc. IRE*, vol. 42, no. 2, pp. 454–457, Feb. 1953.
- [107] H. J. Riblet, "Comment on "Synthesis of symmetrical branch-guide directional coupler"," *IEEE Trans. Microw. Theory and Tech.*, vol. MTT-18, pp. 47–48, Jan. 1970.

-
- [108] R. Levy, "A new class of distributed prototype filters with applications to mixed lumped/distributed component design", *IEEE Trans. Microw. Theory and Tech.*, vol. 18, no. 12, pp.1064-1071, 1970
- [109] M. C. Horton and R. J. Wenzel, "General theory and design of optimum quarter-wave TEM filters," *IEEE Trans Microw. Theory Tech.*, vol. 13, no. 3, pp. 316–327, May. 1965.
- [110] O. P. Gupta, R. J. Wenzel, "Design tables for a class of optimum microwave bandstop filters," *IEEE Trans. Microw. Theory and Tech.*, vol. 18, no. 7, pp.402–404, Jun. 1970.
- [111] H. J. Carlin and W. Kohler, "Direct synthesis of band-pass transmission line structures," *IEEE Trans. Microw. Theory and Tech.*, vol. MTT-13, pp.283-297, May 1965.
- [112] J. Reed and G. J. Wheeler, "A method of analysis of symmetrical four-port networks," *IRE Trans. on Microw. Theory and Tech.* vol. MTT-4, pp. 246–252, Oct. 1956.
- [113] J. –S. Hong and M. J. Lancaster, *Microwave Filters for RF/Microwave Applications*, John Wiley & Sons, Inc., 2001, New York.
- [114] Ansoft HFSS v.9.2.1, Ansoft Corporation, Pittsburgh, PA 15219–1119, USA.
- [115] Advanced Design System (ADS) 2005a, Agilent Technol., Palo Alto, CA, 2005.
- [116] L. Zhu, "Guided-wave characteristics of periodic coplanar waveguides with inductive loading: Unit length transmission parameters," *IEEE Trans. Microw. Theory Tech.*, vol. 51, no. 10, pp. 2133–2138, Oct. 2003.

-
- [117] J.-S. Hong and M. J. Lancaster, "Transmission line filters with advanced filtering characteristics," in *IEEE MTT-S Int. Dig.*, Jun. 2000, pp. 319–322.
- [118] M. Makimoto and S. Yamashita, "Bandpass filters using parallel coupled stripline stepped impedance resonators," *IEEE Trans. Microw. Theory Tech.*, vol. 28, no. 12, pp. 1413–1417, Dec. 1980.
- [119] P. Cai, Z. Ma, X. Guan, Y. Kobayashi, T. Anada, and G. Hagiwara, "Synthesis and realization of novel ultra-wideband bandpass filters using $3/4$ wavelength parallel-coupled line resonators," *Proc. Asia-Pacific Microw. Conf.*, pp. 159–162, Dec. 2006.
- [120] P. Cai, Z. Ma, X. Guan, Y. Kobayashi, T. Anada, and G. Hagiwara, "A novel compact ultra-wideband bandpass filter using a microstrip stepped-impedance four-modes resonator," 2007 *IEEE MTT-S Int. Microwave Sym. Dig.*, WE3E-05, June 2007, pp. 751–754.
- [121] P. Cai, Z. Ma, X. Guan, G. Zheng, T. Anada and G. Hagiwara, "A compact and low-loss sub-millimeter-wave ultra-wideband bandpass filter," *Microw. Opt. Technol. Lett.*, vol.49, no.2, pp. 481–484, 2007.
- [122] P. Cai, Z. Ma, X. Guan, Y. Kobayashi, T. Anada, and G. Hagiwara, "A compact UWB bandpass filter using two-section open-circuited stubs to realize transmission zeros," 2005 *Asia-Pacific Microwave Conf. Dig.*, pp. 3356–3359, Dec. 2005.
- [123] C.-P. Chen, Z. Ma, N. Nagaoka and T. Anada, "Novel compact ultra-wideband bandpass filter employing short circuited stubs with coupled stepped-impedance resonator," *Proc. 2007 Asia-Pacific Microwave Conf.*, pp.1175-1178 Dec. 2007.

-
- [124] Y. -C. Chiou and J.-T. Kuo, "Compact UWB bandpass filter using stub-loaded multiple-mode resonator", *IEEE Microwave and Wireless Components Letters*, Vol.17, No.11, November 2007, pp. 812-812.
- [125] J.-T. Kuo, W.-H. Hsieh, and M. Jiang, "Design of two-stage UIR and SIR bandpass filters with an elliptic function-like response," in *IEEE MTT-S Int. Dig.*, Fort Worth, TX, Jun. 2004, pp. 1609–1612.
- [126] C. Quendo, E. Rius, and C. Person, "Miniaturized lowpass filters with out-of-band improvement by using defected ground plane structure as shunt series stubs," in *34th European Microw. Conf.*, Amsterdam, pp. 495–498, 2004.
- [127] R. Levy, R. V. Snyder, and S. Shin, "Bandstop filters with extended upper passbands," *IEEE Trans. Microw. Theory and Tech.*, vol. 54, no. 6, pp.2503-2515, Jun. 2006.
- [128] Y.-L. Chen and H.-H. Lin, "Novel broadband planar balun using multiple coupled lines," in *2006 IEEE MTT-S Int. Microwave Symp Dig.*, pp. 1571-1574, June 2006.

Author's Publications

Published/Accepted Journal Papers

- [1] **R. Li**, S. Sun and L. Zhu, "Synthesis design of ultra-wideband bandpass filters with designable transmission poles," *IEEE Microwave and Wireless Components Letters*, vol.19, no.3, pp.284-286, May 2009.
- [2] **R. Li**, S. Sun and L. Zhu, "Synthesis design of ultra-wideband bandpass filters with composite series and shunt stubs," *IEEE Transactions on Microwave Theory and Techniques*, vol. 57, no.3, pp. 684-692, March 2009.
- [3] **R. Li**, S. Sun and L. Zhu, "Direct synthesis of transmission line low-pass/high-pass filters with series stubs," *IET Microwaves, Antennas and Propagation*, vol.3, no.4, pp. 654-662, June 2009.
- [4] **R. Li** and L. Zhu, "Ultra-wideband microstrip-slotline bandpass filter with enhanced rejection skirts and widened upper stopband," *Electronics Letters*, vol. 43, no. 24, pp. 1368-1369, November 2007.
- [5] **R. Li**, and L. Zhu, "Ultra-wideband (UWB) bandpass filters with hybrid microstrip/slotline structures," *IEEE Microwave and Wireless Components Letters*, vol. 17, no. 11, pp. 778-780, November 2007.
- [6] **R. Li** and L. Zhu, "Reply to Comment on "Compact UWB bandpass filter using stub-loaded multiple-mode resonator," *IEEE Microwave and Wireless Components Letters*, vol.17, no.11, pp. 812-812, November 2007.

- [7] **R. Li** and L. Zhu, "Compact UWB bandpass filter using stub-loaded multiple-mode resonator," *IEEE Microwave and Wireless Components Letters*, vol. 17, no.1, pp.40-42, January 2007.
- [8] **R. Li** and L. Zhu, "Stopband-improved dual-mode bandpass filter using side-slit patch Resonator," *Microwave and Optical Technology Letters*, vol.49, no.3, pp.717-720, March 2007.

Conference Paper

- [1] S. Sun, **R. Li**, L. Zhu and W. Menzel, "Studies on Synthesis Design of Ultra-Wideband Parallel-Coupled Line Bandpass Filters With Chebyshev Responses," Submitted to *2009 Asia-Pacific Microwave Conference* in May 2009.
- [2] **R. Li** and L. Zhu, "Compact UWB bandpass filter on hybrid microstrip/slotline structure with improved out-of-band performances", *2008 International Conference on Microwave and Millimeter Wave Technology Proceedings*, vol. 1, pp. 6-9, April 2008.