METAMATERIAL-INSPIRED MINIATURIZED MULTI-BAND MICROWAVE FILTERS AND POWER DIVIDERS

by

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Abstract

Metamaterial-Inspired Miniaturized Multi-Band Microwave Filters and Power Dividers

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Integration of more communication standards in one microwave wireless device created a demand on developing compact, low-cost, and robust multi-band microwave components. This dissertation presents three studies for designing miniaturized and multi-band circuits that can be used for multi-band radio frequency (RF) front-ends. These three studies are the design of dual-band and tunable bandpass filters as well as dual- and triple-band equal-split power dividers/combiners. The dual-band filter is based on split ring resonators and double slit complementary split ring resonators. A dual-band prototype three-stage Chebyshev filter, with a fractional bandwidth of 2% at 0.9 GHz and a fractional bandwidth of 3% at 1.3 GHz with equal-ripple of 0.4 dB at both passbands, is presented. The overall size of the dual-band filter is three times smaller compared to edge-coupled microstrip filters. Good out-of-band signal rejection (< 38 dB) and insertion losses (< 4.9 dB for the lower passband and < 2.7 dB for the upper passband) are achieved. The proposed tunable filter is designed from varactor loaded split ring resonators. The size of the tunable filter is reduced by a factor of 3.5 compared to quarter wavelength-based coupled line filters. The power divider is based on composite right- and left-handed transmission lines. Dual-band and triple-band power divider prototypes are designed, fabricated, and tested. The passbands of the triple-band Wilkinson power divider are centered at 0.8 GHz, 1.3 GHz, and 1.85 GHz, and the
passbands of the dual-band Wilkinson power divider are centered at 0.7 GHz, 1.5 GHz. The triple-band divider has a length of 0.66 wavelength in the substrate and its size is reduced to 3/4 of right-handed transmission line-based Wilkinson power dividers. The dual-band power divider has wide fractional bandwidths (\(\sim 20\%\) at the lower passband and \(\sim 41\%\) at the upper passband). Excellent input matchings (input return losses < 29 dB), output matchings (output return losses < 23 dB), and output port isolations (< 24 dB) are achieved at all passbands of the power dividers. The proposed filters and power dividers are compact and low-cost, and are promising candidates for the miniaturization and cost-reduction of multi-band microwave wireless system.
To my parents and to my brothers for their love and continuous support.
## Contents

<table>
<thead>
<tr>
<th>Section</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>Abstract</td>
<td>iii</td>
</tr>
<tr>
<td>List of Tables</td>
<td>viii</td>
</tr>
<tr>
<td>List of Figures</td>
<td>ix</td>
</tr>
<tr>
<td>1 Introduction</td>
<td>1</td>
</tr>
<tr>
<td>1.1 Basic Wireless Transceiver Architecture</td>
<td>1</td>
</tr>
<tr>
<td>1.2 Demands for Wireless Transceivers</td>
<td>3</td>
</tr>
<tr>
<td>1.3 Promise of Metamaterial-Inspired Microwave Circuits</td>
<td>4</td>
</tr>
<tr>
<td>1.4 Dissertation Overview</td>
<td>7</td>
</tr>
<tr>
<td>2 Miniaturized Dual-Passband Microstrip Filter Based on Double-Split Complementary Split Ring and Split Ring Resonators</td>
<td>9</td>
</tr>
<tr>
<td>2.1 Introduction</td>
<td>9</td>
</tr>
<tr>
<td>2.2 Design Methodology</td>
<td>10</td>
</tr>
<tr>
<td>2.3 Results and Discussions</td>
<td>12</td>
</tr>
<tr>
<td>2.3.1 Using CSRR for the Lower Band</td>
<td>13</td>
</tr>
<tr>
<td>2.3.2 Degree of Freedom in Designing Two Passbands</td>
<td>15</td>
</tr>
<tr>
<td>2.4 Conclusion</td>
<td>15</td>
</tr>
<tr>
<td>3 Dual-Bandpass Filters with Individually Controllable Passbands</td>
<td>16</td>
</tr>
<tr>
<td>3.1 Introduction</td>
<td>16</td>
</tr>
<tr>
<td>3.2 Design Methodology</td>
<td>18</td>
</tr>
<tr>
<td>3.2.1 The Basic Cell– A Single Stage Dual-Bandpass Filter</td>
<td>18</td>
</tr>
<tr>
<td>3.2.2 Multi-Stage Dual-Bandpass Filter Implementation</td>
<td>21</td>
</tr>
<tr>
<td>3.3 Design Example and Experimental Results</td>
<td>25</td>
</tr>
<tr>
<td>3.4 Conclusion</td>
<td>32</td>
</tr>
<tr>
<td>4 A Tunable Bandpass Filter Based on Varactor Loaded Split-Ring Resonators</td>
<td>33</td>
</tr>
<tr>
<td>4.1 Introduction</td>
<td>33</td>
</tr>
<tr>
<td>4.2 Design of the Tunable Microstrip Filter</td>
<td>34</td>
</tr>
<tr>
<td>4.2.1 Basic Tunable Filter Module</td>
<td>34</td>
</tr>
<tr>
<td>4.2.2 Multi-Stage Tunable Bandpass Filter Implementation</td>
<td>37</td>
</tr>
<tr>
<td>4.3 Conclusion</td>
<td>39</td>
</tr>
</tbody>
</table>
5 Dual- and Triple-Band Wilkinson Power Dividers Based on Composite Right- and Left-Handed Transmission Lines ........................................... 40
  5.1 Introduction ......................................................... 40
  5.2 Theory and Design Equations ........................................ 42
    5.2.1 CRLH TLs .................................................. 42
    5.2.2 Even-Mode Analysis ........................................ 45
    5.2.3 Odd-Mode Analysis ......................................... 49
  5.3 Implementation and Experimental Results ........................... 53
    5.3.1 Triple-Band Equal-Split Wilkinson Power Divider ............... 54
    5.3.2 Dual-Band Equal-Split Wilkinson Power Divider ................. 57
  5.4 Conclusion ......................................................... 59

6 Summary and Future Work .............................................. 63
  6.1 Summary .......................................................... 63
  6.2 Future Work ...................................................... 64

References ............................................................... 67

Appendix ................................................................. 74

Vita ................................................................. 81
List of Tables

Table | Page
---|---
5.1 | Design parameters of the triple-band equal-split Wilkinson power divider $f_1 = 0.8$ GHz, $f_2 = 1.3$ GHz, and $f_3 = 1.85$ GHz ($N=3$, $k=0.6$). 55
5.2 | Simulated and measured parameters of the triple-band equal-split Wilkinson power divider. 58
5.3 | Design parameters of the dual-band equal-split Wilkinson power divider $f_1 = 0.7$ GHz and $f_2 = 1.5$ GHz ($N=3$, $k=0.5$). 59
5.4 | Simulated and measured parameters of the dual-band equal-split Wilkinson power divider. 62
## List of Figures

<table>
<thead>
<tr>
<th>Figure</th>
<th>Page</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.1</td>
<td>Typical transceiver architecture, (a) receiver and (b) transmitter. Low-noise amplifier (LNA), variable gain amplifier (VGA), phase locked loop (PLL), voltage controlled oscillator (VCO), automated gain control (AGC), power amplifier (PA), analog-to-digital converter (A/D), digital-to-analog converter (D/A).</td>
</tr>
<tr>
<td>1.2</td>
<td>Schematic layout of (a) SRR, (b) CSRR, and (c) DS-CSRR (metal regions are in dark gray).</td>
</tr>
<tr>
<td>1.3</td>
<td>The schematic of a composite right- and left-handed transmission line (CR/LH TL) unit cell (N=1) where right-handed section is implemented by transmission line and left-handed section is implemented by lumped elements.</td>
</tr>
<tr>
<td>1.4</td>
<td>Simulated $S_{21}$ and $S_{11}$ magnitudes of the CR/LH TL with $C_{11}=6$ pF, $Z_{11}=50 \Omega$, $L_{11}=14.1$ nH, $l_{11}=24.78$ mm (red), and $C_{11}=24$ pF, $Z_{11}=25 \Omega$, $L_{11}=14.1$ nH, $l_{11}=24.78$ mm (blue).</td>
</tr>
<tr>
<td>1.5</td>
<td>Simulated $S_{21}$ phase shift of the CR/LH TL with $C_{11}=6$ pF, $Z_{11}=50 \Omega$, $L_{11}=14.1$ nH, $l_{11}=24.78$ mm (red), and $C_{11}=24$ pF, $Z_{11}=25 \Omega$, $L_{11}=14.1$ nH, $l_{11}=24.78$ mm (blue).</td>
</tr>
<tr>
<td>2.1</td>
<td>Layout of the dual-bandpass microstrip filter module.</td>
</tr>
<tr>
<td>2.2</td>
<td>DS-CSRR and the CR on the bottom plane.</td>
</tr>
<tr>
<td>2.3</td>
<td>Equivalent lumped element models of the (a) upper band and the (b) lower band.</td>
</tr>
<tr>
<td>2.4</td>
<td>The dual-band filter basic cell fabricated with Rogers RO3010 substrate.</td>
</tr>
<tr>
<td>2.5</td>
<td>Measured response and Agilent’s Momentum simulation.</td>
</tr>
<tr>
<td>3.1</td>
<td>Schematic view of (a) a rectangular SRR, (b) a rectangular DS-CSRR, (c) the proposed single stage dual-band bandpass module, and (d) its side cut view.</td>
</tr>
<tr>
<td>3.2</td>
<td>Equivalent lumped element circuit model of the basic cell with LC tanks.</td>
</tr>
<tr>
<td>3.3</td>
<td>Simulated scattering parameter results of the basic cell on the substrate RO3010 with 1.27 mm thickness.</td>
</tr>
</tbody>
</table>
3.4 Equivalent circuit model of the proposed three-stage dual-band bandpass filter. 23
3.5 Schematic view of the proposed three-stage dual-band bandpass filter. 24
3.6 External quality factor of SRR as a function of gap-width $s_1$ and the width $w_1$ at the center frequency of upper passband (1.3 GHz) on RO3010 substrate with a thickness of 1.27 mm. 26
3.7 External quality factor of DS-CSRR as a function of gap-width $s_1$ and the width $w_1$ at the center frequency of lower passband (0.9 GHz) on RO3010 substrate with a thickness of 1.27 mm. 27
3.8 Coupling coefficient between the second SRR and first/last SRRs as a function of the width $w_2$ ($s_2=0.1$ mm) at the center frequency of the upper passband (1.3 GHz) on RO3010 substrate with a thickness of 1.27 mm. 28
3.9 Coupling coefficient between the second DS-CSRR and first/last DS-CSRRs as a function of the width $n_2$ ($u_2=3.8$ mm) at the center frequency of the lower passband (0.9 GHz) on RO3010 substrate with a thickness of 1.27 mm. 29
3.10 The simulated and measured scattering parameters of the proposed three-stage dual-band Chebyshev bandpass filter (0.4 dB ripple, 2% and 3% FBWs at 0.9 GHz, and 1.3 GHz, respectively). 30
3.11 Photographs of the fabricated (a) top, (b) middle, and (c) ground layers of the proposed three-stage dual-band Chebyshev bandpass filter on RO3010 substrate ($\varepsilon_r=10.2$, thickness=1.27 mm, and $\tan\delta=0.0035$). 31
4.1 Schematic of the basic tunable filter module. Metal regions are depicted in gray. $w_1=10$ mm, $w_2=2.9$ mm, $w_3=1$ mm, $w_4=0.5$ mm, $w_5=8$ mm, $d_1=1.5$ mm, $d_2=8$ mm. 35
4.2 Simulated $S_{11}$ and $S_{21}$ parameters of the basic tunable filter module. 36
4.3 Measured $S_{11}$ and $S_{21}$ parameters of the basic tunable filter module. 36
4.4 Schematic of the third-order tunable bandpass filter: $d_1=d_4=1$ mm, $d_2=0.85$ mm, $d_3=2.05$ mm, $w_1=1.5$ mm, $w_2=1$ mm, $w_3=12$ mm, $w_4=2.9$ mm. All the resonators are the same size. 37
4.5 Simulated $S_{11}$ and $S_{21}$ parameters of the tunable filter. 38
4.6 Measured $S_{11}$ and $S_{21}$ parameters of the tunable filter. 38
4.7 Fabricated third-order tunable bandpass filter. 39
5.1 Schematic of the proposed Wilkinson power divider. 43
5.2 The schematic of the CRLH TL unit cell \((N=1)\) used in the proposed power divider. .................................................. 43

5.3 The equivalent circuit of the proposed power divider for even-mode analysis. ........................................... 50

5.4 The equivalent circuit of the proposed power divider for odd-mode analysis. ............................................ 50

5.5 Fabricated triple-band equal-split Wilkinson power divider with \(f_1 = 0.8\) GHz, \(f_2 = 1.3\) GHz, and \(f_3 = 1.85\) GHz. ................................................................. 56

5.6 Simulated and measured group delay and \(S_{21}\) (insertion loss) parameters of the triple-band equal-split Wilkinson power divider. ................................................................. 56

5.7 Measured and simulated \(S_{32}\) (output isolation) and \(S_{11}\) (input reflection) parameters of the triple-band equal-split Wilkinson power divider. Black and blue double arrow lines represent measured and simulated bandwidths, respectively. ................................................................. 57

5.8 Measured and simulated output return loss parameters of the triple-band equal-split Wilkinson power divider. ................................................................. 58

5.9 Fabricated dual-band equal-split Wilkinson power divider with \(f_1 = 0.7\) GHz and \(f_2 = 1.5\) GHz. ................................................................. 60

5.10 Simulated and measured group delay and \(S_{21}\) (insertion loss) parameters of the dual-band equal-split Wilkinson power divider. ................................................................. 60

5.11 Measured and simulated \(S_{32}\) (output isolation) and \(S_{11}\) (input reflection) parameters of the dual-band equal-split Wilkinson power divider. Black and blue double arrow lines represent measured and simulated bandwidths, respectively. ................................................................. 61

5.12 Measured and simulated output return loss parameters of the dual-band equal-split Wilkinson power divider. ................................................................. 61
Chapter 1
Introduction

Wireless communication technology has evolved enormously beginning from early 19th century. Many wireless communication standards, i.e., wireless local area networks (WLANs), global positioning system (GPS), and code division multiple access (CDMA), have been developed throughout the world and apparently more standards are to emerge in the near future [1]. The integration of separate standards into one unit increases the size, cost, and complexity of the wireless systems. The need for the design of the low-cost, compact, and robust radio frequency (RF) components operating at multiple frequency bands became apparent in order to meet the demands of next generation wireless systems.

In a wireless system, transmitter-receiver or transceiver is the part of the system which receives and sends the modulated signal over a transmission medium and is one of main contributors of the quality of the transmission [2]. Bit error rate (BER) of the signals available to the baseband of a wireless system or sent to transmission medium is set depending on the signal-to-noise ratio (SNR) performance of the transceiver. Moreover, transceivers occupy considerably larger area and consume more power compared to the rest of the system.

1.1 Basic Wireless Transceiver Architecture

There have been a few different architectures of transceivers such as direct conversion, super-heterodyne, and low-intermediate frequency architectures [3], and the selection of the topology is made based on the required power consumption, performance criterion, and cost. In this section, we present typical receiver (Fig. 1.1(a)) and transmitter (Fig. 1.1(b)) architectures including typical main sections, i.e., frequency synthesizers, antennas, modulators, amplifiers, power divider/splitter, and filters which are commonly used in all topologies. The common components in both sections, i.e., antenna, PLL, VCO, are usually
Fig. 1.1: Typical transceiver architecture, (a) receiver and (b) transmitter. Low-noise amplifier (LNA), variable gain amplifier (VGA), phase locked loop (PLL), voltage controlled oscillator (VCO), automated gain control (AGC), power amplifier (PA), analog-to-digital converter (A/D), digital-to-analog converter (D/A).
shared in order to reduce power and overall size.

The antenna is an interface between the transmission medium and the transceiver. The receiver antenna takes the modulated signal from the transmission medium and passes it through a bandpass filter to cancel unwanted harmonics and to reduce the inference. The received signal is amplified by low-noise amplifier (LNA) and down-converted to baseband for further processing. The selectivity of the bandpass filter and the sensitivity of LNA are main contributors of receiver’s SNR performance. On the transmitter side, the modulated baseband signal is up-converted by mixer and boosted by power amplifier (PA). The mixers are one of the main sources of intermodulation distortions (IMs) and power amplifiers dissipate the most of the available power and occupy considerably large area. All the blocks of the transceiver need to be carefully designed in a wireless system to reduce the size and power consumption.

1.2 Demands for Wireless Transceivers

A direct solution for today’s demand on the design of robust, low-cost, compact multi-standard wireless systems is to miniaturize and to integrate each components in Fig. 1.1 as much as possible without degrading transceiver’s overall performance at any of the standards. Multi-band components are key devices to reduce the size and the cost of a multi-standard wireless system because the overall becomes almost two times smaller compared to the systems implemented using single-band components. That makes multi-band components very attractive for the miniaturization of wireless transceivers. Multi-band transceiver components, such as multi-band bandpass filter [4–6] and dual-band power amplifiers [7–9], have been employed in both transmitters and receivers.

Another solution for the current demand of multi-standard wireless systems is to multi-functional transceiver components. The use of a component, i.e., filter, antenna, amplifier, for each communication standard further increases the volume of the system. Tunable and reconfigurable components are promising alternatives to multi-band components for the design of compact, possibly low-cost and robust multi-standard wireless system.

Although multi-band and tunable microwave transceiver components are beneficial to
reduce the number of components in wireless systems so the size and power consumption, there are many critical performance and design issues of multi-band and tunable components to solve compared to single-band RF transceiver components. For this reason, there is an ongoing demand for higher level miniaturization of individual single-band transceiver components.

1.3 Promise of Metamaterial-Inspired Microwave Circuits

Among various methods to scale down the size of wireless systems, the use of recently developed metamaterial-inspired structures has been showing continued promise. Metamaterials also called left-handed materials (LHMs), which is first introduced by Verselogo [10], have created a new research in the area of microwave circuit design. The unique properties of metamaterials (having negative permeability and permittivity) have allowed development of novel applications and devices. Metamaterials are implemented with split ring resonators (SRRs) as proposed by Smith [11] (Fig. 1.2(a)), complementary split ring resonators (CSRRs) [12] (Fig 1.2(b)), and double slit complementary split ring resonators (DS-CSRRs) [13] (Fig. 1.2(c)). SRRs, CSRRs, and DS-CSRRs are high-Q resonators and can be excited by a time-varying electric or magnetic field. The promise of SRRs and CSRRs is to reduce circuit dimensions due to fact that these resonators can be designed with dimensions much smaller than signal wavelength at resonance. A bandpass filter with controllable bandwidth using CSRRs, capacitive gaps, and shorted inductive lines has been designed [14]. Also, a band-reject filter using CSRR is presented by Falcone [13]. A varactor loaded band-reject tunable filter is designed with SRRs [15]. These works have shown that filters constructed from SRRs and DS-CSRRs result in compact sizes.

Another way of implementing planar metamaterials, namely composite right- and left-handed transmission line (CR/LH TL), is shown in Fig. 1.3 [16]. CR/LH TLs have nonlinear phase slopes. The simulated scattering parameters of two different designs of CR/LH TLs are shown in Fig. 1.4 and Fig. 1.5. As it can be seen from Fig. 1.4 and Fig. 1.5, the phase slope and phase delay at a particular frequency can be arbitrarily designed. The values of CR/LH TL elements can be adjusted to have two desired impedances at
two specific frequencies. Due to these interesting properties, a CR/LH TL can be shorter than a conventional right-handed (RH) transmission line, so the dimensions of multi-band microwave devices can be reduced using these novel transmission lines.

CR/LH TLs and SRRs have been extensively used in design of both active and passive microwave devices such as amplifiers, filters, antennas, couplers, and phase shifters. Although both implementations (SRR and CR/LH TL) have reported advance in enhancing and miniaturization of microwave wireless communication systems, the CR/LH TL approach is proven to have built-in properties of being wideband and low-loss [17] and possibly have wider applications. A class-F power amplifier using a harmonic tuner based on CR/LH TLs is proposed by Dupuy [18]. The tuning circuit design proposed in this paper improves the power added efficiency (PAE) of the amplifier. Also, a dual-band class-E
Fig. 1.4: Simulated $S_{21}$ and $S_{11}$ magnitudes of the CR/LH TL with $C_{11}=6 \text{ pF}$, $Z_{11}=50 \Omega$, $L_{11}=14.1 \text{ nH}$, $l_{11}=24.78 \text{ mm}$ (red), and $C_{11}=24 \text{ pF}$, $Z_{11}=25 \Omega$, $L_{11}=14.1 \text{ nH}$, $l_{11}=24.78 \text{ mm}$ (blue).

Fig. 1.5: Simulated $S_{21}$ phase shift of the CR/LH TL with $C_{11}=6 \text{ pF}$, $Z_{11}=50 \Omega$, $L_{11}=14.1 \text{ nH}$, $l_{11}=24.78 \text{ mm}$ (red), and $C_{11}=24 \text{ pF}$, $Z_{11}=25 \Omega$, $L_{11}=14.1 \text{ nH}$, $l_{11}=24.78 \text{ mm}$ (blue).
power amplifier using CR/LH TLs has been presented by Seung [19]. The design of dual-band amplifiers is a difficult task since matching networks at input and output ports need to operate at two desired frequencies. A wideband bandpass filter using CR/LH TLs is designed by Gil [20]. This design uses complementary split rings resonators (CSRRs) and capacitive gaps edged on the top plane of the substrate. A dual-band filter based on CR/LH TLs where the passbands of the filter does not necessarily have to be at first and third harmonics is designed [21]. A multiplier with low-noise figure is designed using defected ground plane and CR/LH TLs by Song [22]. A CR/LH TL-based coupled line directional coupler with arbitrary bandwidth and coupling is proposed by Caloz [23]. Using classical balanced mixer topology, a dual-band mixer integrated with CR/LH TLs to reduce the circuit size is presented by Paco [24]. A 180 degree broadband phase shifter using nonlinear phase property of CR/LH TLs is designed [25]. Although many studies have been done so far, there still exist tremendous challenges and many aspects to discover. Microstrip filters employing SRRs are mainly single-band and furthermore, there is still a vast need to research on miniaturization of multi-band transceiver components using CR/LH TLs.

1.4 Dissertation Overview

This dissertation is focused on miniaturization of multi-band or tunable microwave transceiver components. Novel dual-band and tunable bandpass filters as well as dual- and triple-band power dividers are proposed, designed, fabricated, and measured for demonstration. This dissertation is prepared in multiple-paper format.

Chapter 2 is the first paper, which describes a compact dual-band microstrip bandpass filter module.

Chapter 3 is the second paper, which demonstrates the miniaturization of dual-band bandpass microstrip filters. The control over each passband characteristics of the proposed dual-band bandpass filter is demonstrated with the design of a dual-band Chebyshev bandpass filter.

Chapter 4 is the third paper and presents a compact tunable bandpass filter based on reverse-biased varactor diode loaded SRRs. The scattering parameter performances of the
proposed single tunable block and coupled resonator bandpass filter implementations are presented. Both the single block and the coupled resonator filter are fabricated and tested to verify the simulations.

Chapter 5 is the fourth paper and presents dual- and triple-band Wilkinson power dividers based on CR/LH TLs. The proposed Wilkinson power dividers have compact sizes as well as wide fractional bandwidths compared to the divider designs implemented by conventional right-handed microstrip transmission lines.

Chapter 6 draws some conclusions and this chapter is concluded with ideas about future work.
Chapter 2

Miniaturized Dual-Passband Microstrip Filter Based on Double-Split Complementary Split Ring and Split Ring Resonators

Abstract

This paper presents a miniaturized dual-passband filter module designed using a double-split complementary split ring resonator (DS-CSRR) and a split ring resonator (SRR). The use of SRR results in a significant size reduction of the filter comparing with coupled-line filters. Two passbands are individually printed on two sides of a Rogers 3010 substrate, consequently providing a novel and compact integration. Coupling between two bands is weak, so they can be independently designed and tuned. Both bands operate at fundamental mode, providing an increased stability. A prototype dual-band filter basic cell is fabricated and the measurement agrees well with simulations by Agilents Momentum.

2.1 Introduction

The use of ever-broadening communication capacities illuminates the importance of multi-band antennas and RF front-ends. With the rapid increase in communication capacity and new functions such as GPS and Bluetooth, it is fair to expect that all handsets will become compatible with multi-bands in the near future, and consequently require an efficient integration of multi-band devices. Besides integration, circuit miniaturization is another goal for multi-band front-ends. Dual-passband filters have been reported in response to these challenges [4–6]. However, these designs either lack control over the bandwidths

of each passband [4], or have relatively large circuit size [5, 6]. This paper presents a miniaturized dual-passband microstrip filter module that can serve as building block for higher order filter implementation. The design is based on planar microstrip technology with the advantage of being robust and easy to integrate. Two passbands can be tuned independently within a large frequency range. The dual-band filter has a potential use in integrating multiple bands such as two GSM bands or GSM and Bluetooth in one unit.

The basic cell presented is an integration of planar microstrip double-split complimentary split ring resonator (DS-CSRR) and split ring resonator (SRR) similar to those reported by Marques [13]. A microstrip SRR is two concentric planar rings with splits printed on a thin dielectric substrate, and is a planar version of SRRs [11, 26, 27]. Because it operates at a quasi-TEM mode, a SRR can resonate with a size much smaller than conventional microwave resonators. Garcia-Lampererez et al. designed a dual-band filter with SSRs [28], however, the second passband of Garcia-Lampererezs filter is actually due to the higher order resonance of the SRR, and hence lacks the freedom in designing two independent passband. Also, stability can be another issue, since the resonance is not from the fundamental mode.

A complementary split ring resonator (CSRR) [12] is the negative image of a SRR etched on the ground plane of a FR-4 type of substrate, and a DS-CSRR is a CRSS with two extra slits [13]. Bonache et al. presented a design with a CSRR as the basic cell [14]. Given the resonant nature of SRR and DS-CSRR (CSRR), and the fact that they can be fabricated on opposite sides of a circuit board, it is intuitive to design a compact dual band filter with SRR and DS-CSRR, where SRR and DS-CSRR represents two passbands, respectively. Designing a filter with split ring type resonators results in a size reduction, and integrating two bands on each side of the same board offers a further miniaturization. It is the purpose of this work to show a novel, compact, dual-passband filter architecture with flexibility in the passband design and increased stability.

2.2 Design Methodology

The circuit layout of the dual-band microstrip filter module is shown in Fig. 2.1. A SRR (one with dark gray color in the picture) is printed on the top plane of a high-frequency
laminate and a DS-CSRR etched on bottom (ground) plane is represented by the light gray color. The bottom plane is shown in more detail in Fig. 2.2. An extra complementary ring (CR, a negative image of a ring, denoted by the inner radius $R_1$ and outer radius $R_2$ in Fig. 2.2) inside the DS-CSRR is etched on the ground plane to provide the ground for SRR without disturbing the resonance of DS-CRSS. With this layout, the ground for SRR can be achieved by grounding the metal disk (the one with the radius of $R_1$ in Fig. 2.2). We achieved the grounding by making use of four grounding patches as shown with the dashed lines and marked by GP in Fig. 2.1. The SRR and DS-CSRR correspond to the upper passband and lower passband, respectively. Both resonators share the same feed-lines for excitation. The SRR is excited by the axial magnetic field coupled from a quarter wave arch connected to the 50 Ohm feed-line. The DS-CSRR is excited by the electric field coupled from the CR, which couples with the signal from the feed line on the top plane.

We found that the coupling between two resonators is weak and accordingly the lumped element model is derived as shown in Fig. 2.3. Figure 2.3(a) is the lumped element model for
Fig. 2.2: DS-CSRR and the CR on the bottom plane.

the upper band with SRR as its resonator and Fig. 2.3(b) is for the lower band with DS-SRR as the resonator. The model is derived from low-pass filter prototype by element/frequency transformations [29], and modeling SRR and DS-CSRR with a simple LC circuit is in line with the previously reported developments [27,30] that the simplification is due to the small electric size of split ring type of resonators. $L_r$ and $C_r$ in Fig. 2.3 are equivalent lumped inductance and capacitance for DS-CSRR and $L_s$ and $C_s$ are for SRR. $L$ is the inductance between the feed port and the arch, $C_g$ is the gap capacitance between arch and SRR, and $C_c$ is capacitance between the arch shape feed-line and the DS-CSRR.

2.3 Results and Discussions

We designed a dual-passband filter with center frequencies at 900 MHz and 1.7 GHz, and fabricated the filter (Fig. 2.4) with Rogers RO3010 substrate, which has relative dielectric constant 10.2 and thickness 1.27 mm. Four copper patches (Fig. 2.4(c)) are used to ground the center disk. The measurements were performed with HP 8510 network analyzer. Simulated results with Agilent’s Momentum and measurements are plotted in Fig. 2.5. The stop-band signal level between lower and upper bands is about -20 dB which is insufficient
Fig. 2.3: Equivalent lumped element models of the (a) upper band and the (b) lower band. if a single module is used as a filter by itself. However, once the module is cascaded to implement higher order filter, the level would reduce down to -40∼-50 dB. Cascading the presented basic cell to produce a higher order filter response is the further implementation of the topology. The measurements agree well with the simulations, and the small shift in the center frequency is mainly due to the parasitic coupling through the grounding patches. With increased fabrication precision, the shift can be reduced. The overall size of the filter is 27x27 mm$^2$, and is about three times smaller than a conventional half wavelength resonator-based bandpass filter. We also gain an additional scaling factor of two by integration two resonators on one board. Therefore, the overall size miniaturization factor is larger than five.

2.3.1 Using CSRR for the Lower Band

It is possible to choose a CSRR for the lower band and the design process is the same to the one presented. The choice of DS-CSRR and CSRR depends on the difference in center frequencies of the two passbands ($f_u$ and $f_l$). When $f_u \geq 1.5 f_l$, DS-CSRR is chosen to achieve sufficient coupling from the feed-lines on the top plane. Otherwise, when $f_u$ and
Fig. 2.4: The dual-band filter basic cell fabricated with Rogers RO3010 substrate.

Fig. 2.5: Measured response and Agilent’s Momentum simulation.
$f_l$ are closer to each other, a CSRR can be chosen to obtain a smaller circuit layout than by using a DS-CSRR.

2.3.2 Degree of Freedom in Designing Two Passbands

The second order mode (the one higher than the fundamental mode) resonance of the DS-CSRR occurs at about $2.2 f_l$. Therefore, in order not to corrupt the frequency response of the upper band, one needs to design the second band such that $f_u \leq 2f_l$.

2.4 Conclusion

A new method of integrating two passbands into one circuit board is demonstrated. The proposed filter is a basic cell and can be cascaded to implement periodic bandpass filters such as Chebyshev bandpass filter. In order to realize admittance inverters between the modules, the coupling capacitances $C_c$ and $C_g$ can be designed to have 90 degree phase shift at lower and upper band, respectively. Two passbands can be independently designed within the limit discussed in sec. 2.3.2. The design provides an alternative method for current integration techniques and has potential applications in miniaturizing of the multiband RF frontends.
Chapter 3
Dual-Bandpass Filters with Individually Controllable Passbands

Abstract

This paper presents a novel dual bandpass filter based on split ring resonators (SRRs) and double slit complementary split ring resonators (DS-CSRRs). The size of the filter is small and two passbands can be individually designed. The basic cell of the filter is presented and analyzed, and then the multi-stage dual passband filter is achieved by cascading the basic cell. The design graphs for external quality factors of the resonators at input and output stages and the coupling coefficient between the adjacent resonators are constructed. The design graphs are utilized to determine proper geometric parameters of each filter stage for a given filter specification. As an example, a prototype three-stage Chebyshev filter with a fractional bandwidth of 2% at 0.9 GHz and a fractional bandwidth of 3% at 1.3 GHz is demonstrated. The prototyped filter has an equal-ripple of 0.4 dB at both passbands. The measurements of the prototyped filter agree well with simulation results. The center frequencies and the fractional bandwidths of two passbands can be individually designed with a large degree of flexibility compared to dual-band filters that utilizes resonances of higher order modes. The proposed design integrates two passbands on the two sides of the same substrate and the overall size reduction can be as high as a factor of three compared to edge-coupled microstrip filters.

3.1 Introduction

Rapid developments in wireless microwave communication systems have created needs for multi-band operations. It is favored to integrate multi-standard operations such as global system for mobile communications (GSM), code-division multiple-access (CDMA)
and industrial-scientific-medical (ISM) in one unit. To meet such a demand of multi-band integration, dual-band microwave components, i.e., antennas [31, 32], rectennas [33], couplers [34, 35], and bandpass filters [4, 36–41] have been developed. This paper is aimed to design an improved dual-band filter because filters are perhaps one of the most used components in wireless communication. Although there are a large number of dual-band filters, their sizes are relatively large [36–39]. A dual-bandpass filter based on dual-feeding structure was presented [40], but this design suffers from very low attenuation between the passbands. Dual bandpass filters using stepped-impedance resonators were developed by properly selecting the relevant impedance or strip width ratio [4, 41]. The resonant frequencies of the stepped impedance resonators are dependent, and therefore it is difficult to simultaneously achieve two passbands with adjustable fractional bandwidths (FBWs).

On the other hand, split ring resonators (SRRs) [11], double split SRRs and their complements – complementary split ring resonators (CSRRs) and double slit CSRRs (DS-CSRRs) – have been used to design miniaturized filters due to fact that sizes of these resonators are much smaller than their wavelengths at resonant frequencies [13]. SRRs have been used to improve the out-of-band performance of bandpass filters [42, 43]. Compact single-band bandpass filters with controllable bandwidths based on SRRs and CSRRs have been presented [14, 44]. A dual-bandpass filter using SRRs has been reported [28] but this design lacks flexibility in designing the bandwidth.

In this paper, we present a novel compact dual bandpass filter implementation based on SRRs and DS-CSRRs. A single-stage dual-band filter module as the basic cell is analyzed, and then is cascaded to multi-stage filters. The coupling between the resonators of adjacent stages and the external quality factors of SRR and DS-CSRR in each stage can be controlled by adjusting the basic cell in each stage. The bandwidth and passband characteristics of each passband can be individually designed. The design methodology for multi-stage implementation of standard filter approximations is presented. As an example, a three-stage Chebyshev dual-band filter is designed and prototyped. The simulated and measured results agree well. The proposed design is compact and easy to be fabricated, suggesting
promises to be implemented in multi-frequency communication systems.

3.2 Design Methodology

3.2.1 The Basic Cell– A Single Stage Dual-Bandpass Filter

The basic cell of the proposed filter contains a rectangular SRR and a rectangular DS-CSRR constructed from concentric rectangular rings shown in Fig. 3.1(a) and (b), respectively. A similar structure was presented in Chapter 2 and this paper presents a more complex design as well as an improved grounding method for the DS-CSRR. The schematic and side view of the basic cell is presented in Fig. 3.1(c) and (d). The module consists of three copper layers (top, middle, and bottom layers). The SRR is printed on the top layer and the DS-CSRR is etched on the middle layer. The bottom layer is the ground.

RO3010 high-frequency laminates ($\varepsilon_r=10.2$, $h_1=h_3=1.27$ mm and $\tan\delta=0.0035$) are chosen as the substrates. The upper and lower substrates (Fig. 3.1(d)) are laminated with two layers of GORE speedboard C prepreg boards ($\varepsilon_r=2.6$, $h_2=2\times0.051$ mm and $\tan\delta=0.004$). The input/output (I/O) lines are designed as 1.1 mm to have the characteristic impedance of 50 Ohm on the substrate.

Each RO3010 board is processed separately before laminating them using speedboards. The fabrication steps of the filter are summarized as follows. First, the vias in the lower RO3010 substrate (between bottom and middle layers of the filter module) are drilled and filled with DuPont CB100 conductive paste (volume resistivity=0.00016 $\Omega$/cm). Then the DS-CSRR is etched on the top layer of the lower RO3010 substrate. The bottom copper layer of the upper RO3010 substrate is completely etched off and the SRR is fabricated on the top copper of the upper RO3010 substrate. During the lamination, the speedboards are placed in between the upper and lower RO3010 substrates. The assembly (two substrates and speedboards in between) is sandwiched between a Lauffer 88 ton vacuum lamination press. The press is heated up to 425$^0$ F and a pressure of 370 psi is applied. The assembly is removed from the press after a cycle of 130 minutes and is cooled to room temperature.

The equivalent circuit of the basic cell is presented in Fig. 3.2 where the SRR and
Fig. 3.1: Schematic view of (a) a rectangular SRR, (b) a rectangular DS-CSRR, (c) the proposed single stage dual-band bandpass module, and (d) its side cut view.
DS-CSRR are modeled by LC tanks [30]. The SRR and DS-CSRR are magnetically and electrically coupled to the feed-line, respectively. The resonant frequencies of SRR and DS-CSRR are given by

\[ f_{SRR} = \frac{1}{2\pi\sqrt{L_1C_1}}, \quad \text{and} \quad f_{DS-CSRR} = \frac{1}{2\pi\sqrt{L_2C_2}}, \] (3.1)

where \( L_1, L_2 \) are the equivalent inductances and \( C_1, C_2 \) are the equivalent capacitances of the outer rings of the SRR and DS-CSRR. The values of \( L_1, C_1 \), and hence the resonant frequency of the SRR, is dependent of the physical size \( k \) as well as the split-width (\( g \)), gap-distance (\( s \)), ring-width (\( c \)), and the gap-width between the concentric split rings of the SRR (Fig. 3.1). Increasing \( g, s, c, \) and \( t \) decreases the resonance frequency of the SRR. On the other hand, the dimensions of \( z, u, n, v, y \) determine the values of \( L_2, C_2 \), and hence the resonant frequency of the DS-CSRR.

The external quality factors of the SRR \( (Q_{e1}) \) and DS-CSRR \( (Q_{e2}) \) resonators are determined from

\[ Q_{e1} = Z_0\sqrt{\frac{C_1}{L_1}}, \quad \text{and} \quad Q_{e2} = Z_0\sqrt{\frac{C_2}{L_2}}, \] (3.2)
where $Z_0$ is the characteristic impedance at input and output ports. The bandwidth of the SRR is significantly dependent on the coupling between the input/output line and the SRR, and the ratio $\sqrt{C/L}$ [45]. Decreasing the gap-width ($s$) or increasing length ($w$) increases the coupling, and accordingly increases the resonant bandwidth of the SRR. Similarly, the bandwidth of the DS-CSRR is proportional to the coupling between the DS-CSRR and the transmission line. The coupling between the DS-CSRR and the transmission line can be adjusted by $u$ and $n$. In addition, increasing the slit width ($v$) degrades the quality factor of the DS-CSRR, and results in higher bandwidth.

The simulated frequency response of the basic cell from Agilent’s Momentum is shown in Fig. 3.3. The dimensions of the module are: $k=10.9$, $g=1.6$, $s=0.1$, $l=9.75$, $z=15.5$, $t=7.5$, $m=1.65$, $u=2$, $n=1$, $y=b=0.5$, $w=4$, $a=0.5$, $v=1$ (all in mm). The radius of the vias (total of 28 vias in one cell) is 0.5 mm. The external quality factors of the SRR and the DS-CSRR are calculated to be 38.31 and 75.53. The lumped element values of LC tank for each passband are calculated to be $L_1=0.139$ nH, $L_2=0.103$ nH, $C_1=81.6$ pF, and $C_2=233$ pF. The simulated results show that one can achieve two passband frequencies with controllable external quality factors, which are necessary for multi-stage implementations with standard filter approximations. The rejection level between the passbands in the filter cell is insufficient (around 20 dB), and would improve when cascading the basic cells to multi-stage filters.

It is noted that the coupling between SRR and DS-CSRR is weak, therefore allows one to design two resonators independently. It is also noted that if one intends to utilize only the fundamental mode resonances of SRR and DS-CSRR, the two passbands can be flexibly designed as long as the center frequency of the upper band is less than twice of the center frequency of the lower band.

3.2.2 Multi-Stage Dual-Bandpass Filter Implementation

The basic cell presented in sec. 3.2.1 is the building block for multi-stage dual bandpass filters. The coupled resonator filter synthesis method explained by Hong and Lancaster [29] is used for each passband individually and the design parameters (coupling coefficients,
The coupling between SRRs is achieved from the magnetic coupling through capacitive gaps ($s_1, s_2, s_3$) and the transmission lines ($l_2, l_2'$) (Fig. 3.5). The coupling between the DS-CSRRs is achieved from the electric coupling through the substrate and transmission lines.

The external quality factors of the input and output resonators (i.e., the resonators in the first and the last cells in Fig. 3.5) and the coupling coefficient between the adjacent resonators have to be properly designed in order to synthesize a bandpass filter with standard filter approximations (i.e., Chebyshev, Butterworth, and elliptic responses). The required external quality factors and coupling coefficients of each passband for a given filter
Fig. 3.4: Equivalent circuit model of the proposed three-stage dual-band bandpass filter.

specification can be calculated using the following formulas:

\[
Q_{e11} = Q_{e13} = \frac{g_0g_1}{FBW_1}, Q_{e21} = Q_{e23} = \frac{g_0g_1}{FBW_2},
\]

\[
M_{12} = M_{13} = \frac{FBW_1}{g_1g_2}, \text{ and } M_{22} = M_{23} = \frac{FBW_2}{g_1g_2},
\]

where \(FBW_1\) and \(FBW_2\) are the fractional bandwidths of lower and upper passbands of the bandpass filter, respectively, and \(g_0, g_1, \ldots, g_n\) are the element values of the low-pass filter model.

Using the full-wave Agilent’s Momentum simulations, the coupling between a pair of adjacent SRRs or DS-CSRRs can be extracted from

\[
M = \frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2},
\]

where \(f_{p1}\) and \(f_{p2}\) are \((f_{p2} > f_{p1})\) the two resonance frequencies of the SRRs or DS-CSRRs [29]. It should be noted that the realizable external quality factors and coupling coefficients are limited by the precision of the fabrication process even though theoretically the quality factors and coefficients can be arbitrary values.
Fig. 3.5: Schematic view of the proposed three-stage dual-band bandpass filter.
The external quality factors of the resonators in the first and last stages can be evaluated from the full-wave simulations as

\[ Q_{e11} = Q_{e13} = \frac{f_1}{f^h_1 - f^l_1}, \quad \text{and} \quad Q_{e21} = Q_{e23} = \frac{f_2}{f^h_2 - f^l_2}, \]

where \( f^h_1, f^l_1 \) are the upper and lower 3-dB cut-off frequencies of the lower passband, and \( f^h_2, f^l_2 \) are the cut-off frequencies of the upper passbands.

For a given passband center frequencies and fractional bandwidths, the design procedure of the proposed dual-band filter can be summarized as follows.

1) Determine the dimensions of the SRRs and DS-CSRRs using (3.1).

2) Generate the design graphs using full-wave simulations for the external quality factors, which are functions of \( s, w, u, \) and \( n \) of SRR and DS-CSRR.

3) Determine the parameters \( s, w, u, \) and \( n \) of the first and last stages from the design graphs and (3.3).

4) Use full-wave simulations and (3.6) to generate the coupling coefficients between the first and middle stages.

5) Determine the values for \( s, w, u, \) and \( n \) of the second stage using the coupling coefficient graphs from the step 5) and (3.4).

6) Adjust values of \( g \) and \( v \) for each stage to tune the resonant frequencies to \( f_1 \) and \( f_2 \).

In step 1), one generates initial values for \( s, w, u, \) and \( n \). During step 3) and 5), these values are changed to obtain the proper quality factors and coupling coefficients. The change will shift the resonant frequencies of SRR and DS-CSRR. Therefore, it is necessary to perform step 6) to adjust the passband frequencies back to the \( f_1 \) and \( f_2 \).

**3.3 Design Example and Experimental Results**

As a demonstration, a three-stage dual-band Chebyshev filter with passbands centered at 0.9 GHz and 1.3 GHz is prototyped. The FBWs of the lower and upper passbands are
chosen as 2% and 3%, respectively. Each passband has an equal-ripple of 0.4 dB (i.e., return loss \( \leq -10.56 \) dB). The low-pass filter prototype element values are obtained from the tables provided by Hong and Lancester [29] as \( g_0 = g_4 = 1, \ g_1 = g_3 = 1.4904, \) and \( g_2 = 1.1181. \) The required external quality factors of resonators and coupling coefficients are calculated to be \( Q_{11} = Q_{13} = 74.52, \ M_{12} = M_{13} = 0.0155 \) (DS-SRRs), and \( Q_{21} = Q_{23} = 49.68, \ M_{22} = M_{23} = 0.0232 \) (SRRs). Full-wave Momentum simulations were performed and design graphs for coupling coefficients and resonator external quality factors are extracted for both passbands.

Following the design procedure outlined in sec. 3.2.2, the dimensions of SRRs and DS-CSRRs were determined such that they resonated in the vicinity of 0.9 GHz and 1.3 GHz, respectively. Then we generated the resonator design graph for the quality factor of the SRRs at 1.3 GHz (Fig. 3.6). A wide variety of resonator external quality factors (i.e., from 400 to 22 that corresponds to a resonator bandwidth between 0.25% and 4.54%) can be achieved by varying \( s_1 \) and \( w_1. \) The external quality factor of the SRR is proportional to the magnitude of the coupling between it and the transmission line. Therefore, an increase in \( w_1 \) or a decrease in \( s_1 \) results in a degradation of the quality factor. The gap-width \( g_
Fig. 3.7: External quality factor of DS-CSRR as a function of gap-width $s_1$ and the width $w_1$ at the center frequency of lower passband (0.9 GHz) on RO3010 substrate with a thickness of 1.27 mm.

(Fig. 3.1(c)) of SRRs is adjusted to tune the resonance back to 1.3 GHz. The external quality factor of the DS-CSRR at 0.9 GHz as a function of $t_1$ and $u_1$ is presented in Fig. 3.7. Figure 3.6 and Fig. 3.7 can be also used to determine the dimensions of the last stage. By adjusting $t_1$ and $u_1$, one can achieve an external quality factor ranging between 58 and 272 (corresponds to a resonator bandwidth from 0.37% to 1.72%). Similar to SRRs, $v$ (Fig. 3.1(c)) has to be adjusted to tune the DS-CSRR to resonate back at 0.9 GHz.

The values for $a$, $b$, $c$, $d$ (Fig. 3.1(c)) are chosen as 0.5 mm. From Fig. 3.6 and Fig. 3.7, the values for $s$, $w$, $n$, and $u$ are determined as $s_1 = s_3 = 0.1$ mm, $w_1 = w_3 = 7$ mm, $n_1 = n_3 = 3.25$ mm, and $u_1 = u_3 = 3.8$ mm. The remaining parameters are found to be $k_1 = k_3 = 12.3$ mm, $g_1 = g_3 = 1.68$ mm, $m_1 = m_3 = 2.25$ mm, $t_1 = t_3 = 8.9$ mm, $v_1 = v_3 = 2.8$ mm, $z_1 = z_3 = 17.8$ mm, and $l_1 = l_3 = 12.8$ mm.

The extracted coupling coefficients between the SRRs in the first and middle stage are plotted against $w_2$ with fixed $s_1$, $w_1$, and $s_2$ in Fig. 3.8. The same figure can also be used to determine the coupling between the middle stage and the last stage. Figure 3.9 shows the coupling coefficients for DS-CSRRs as a function of $n_2$ with fixed $n_1$, $u_1$, and $u_2$. The values of $s_2$ and $u_2$ are set to 0.1 mm and 3.8 mm, respectively. It can be seen from Fig. 3.8
and Fig. 3.9 that the coupling between resonators is almost linearly proportional to $w_2$ and $n_2$. The required dimensions for the calculated coupling coefficients are obtained as $w_2=7.5$ mm and $n_2=2.65$ mm from Fig. 3.8 and Fig. 3.9. In the last step of tuning the DS CSRRs and the SRRs back to passband frequencies, the rest of the parameters were determined to be $k_2=12.3$ mm, $g_2=1.62$ mm, $m_2=2.25$ mm, $t_2=8.9$ mm, $v_2=2.24$ mm, $z_2=17.8$ mm, and $l_2=14.1$ mm.

It should be noted that the design graphs are specific to the passband frequency and substrate. When a different substrate or different passband frequency is required, one has to reconstruct the design graphs.

The simulated and measured scattering parameters of the filter are presented in Fig. 3.10. Measurements were performed using an Agilent 8510C vector network analyzer. The dielectric loss, conductivity of copper ($\sigma=5.8e7$ S/m) and the thickness of copper (35 $\mu$m) are included in simulations. As shown in Fig. 3.10, the measured passband and out-of-band performance show good agreement with the simulated results, verifying the design methodology for multi-stage dual-band bandpass filter. Fig. 3.11 shows the photograph of
Fig. 3.9: Coupling coefficient between the second DS-CSRR and first/last DS-CSRRs as a function of the width $n_2$ ($u_2=3.8$ mm) at the center frequency of the lower passband (0.9 GHz) on RO3010 substrate with a thickness of 1.27 mm.

the fabricated filter. The size of this filter is about 0.628 $\lambda_g$ by 0.174 $\lambda_g$, where $\lambda_g$ (extracted from the full-wave simulation) is the guided wavelength on the substrate at the center of the lower passband.

The measured passband center frequencies are 875 MHz and 1.265 GHz and are slightly shifted from the simulation. The shifts are mainly due to fabrication. The widths of lines and slots are in the vicinity of 2 mm and small variation due to fabrication inaccuracy may result in a change in frequency. When assembling two bands together, heating and pressure are applied. Although the material properties of Rogers substrate may not vary during the process, the thickness of the GORE speedboards may have been affected. Also, the heating and pressing can result in non-uniform distribution of the dielectric constant of the speedboard. All of these factors can contribute to the shift in the frequency. The measured out-of-band performance is good and shows a close agreement with the simulations. The measured FBWs for the lower and upper passbands are 2.86% and 3.32% (corresponds to 25 MHz and 42 MHz bandwidth), respectively. The measured bandwidths are slightly wider than simulations, which might be attributed to stronger couplings between the stages and
higher external quality factors due to the material and fabrication tolerances. The measured and simulated in-band return losses for both passbands show good agreement and they are higher than 10 dB.

The simulated insertion losses are 3.3 dB and 2.2 dB, respectively, for the lower and upper passband, while the measured insertion losses are seen to be 4.9 dB and 2.7 dB. The difference in the insertion losses can be due to the effect of SMA connectors, mismatches at the I/O ports, and additional radiation losses. Also, it is expected that lower band will suffer from higher insertion loss because there are losses from two substrates for the lower band. It is also found that while the flexibility of designing two passband frequencies in the single-stage filter module is within the range of 2.0 (the center frequency of the upper band is within 2.0 times of the lower band frequency), the three-stage filter shows a flexible range of 1.7. Within such range, one can design two bands independently and the resonators in each band operate at their fundamental modes. Finally, the overall size reduction is a factor of three compared to an edge-coupled microstrip filter and it is possible to have
Fig. 3.11: Photographs of the fabricated (a) top, (b) middle, and (c) ground layers of the proposed three-stage dual-band Chebyshev bandpass filter on RO3010 substrate ($\varepsilon_r=10.2$, thickness=1.27 mm, and tan $\delta=0.0035$).
further reduction because the length of $l_2, l'_2$ (Fig. 3.5) can be easily reduced.

3.4 Conclusion

The paper presents a novel compact dual-band microstrip filter based on SRRs and DS-CSRRs. The characteristics of each passband (i.e., center frequency, FBW, passband ripple level) can be individually and flexibly designed. The design method is validated by demonstrating a three-stage Chebyshev filter with 2% FBW at 900 MHz and 3% FBW at 1.3 GHz. Good in-band and out-of-band performance is achieved. Simulations and measured results are presented and show good agreements. After considering the overall size for each band and the integration of two bands on the same multi-layered circuit board, the size of the demonstrated dual-band filter is three times smaller than an edge-couple microstrip filter and can be further miniaturized. The proposed dual-band filter offers design flexibility and a miniaturized size, and can be potentially adapted for multi-band microwave communication systems.
Chapter 4

A Tunable Bandpass Filter Based on Varactor Loaded Split-Ring Resonators

Abstract

This paper presents an electronically tunable varactor loaded microstrip bandpass filter based on the planar split ring resonators (SRRs). The varactor is a reverse-biased semiconductor diode, and is connected between the concentric rings of the SRR. An individual varactor loaded SRR-based bandpass tunable filter module is analyzed. Then a third order tunable filter with 2% fractional bandwidth and a tuning range from 2.11 GHz to 2.34 GHz is assembled from the basic filter module. A prototype filter is designed, fabricated and tested. The maximum tuning range of the prototype filter is 10.2% and the size of the filter is reduced by a factor of 3.5 compared to coupled line filters.

4.1 Introduction

Frequency tunable filters are of great interest in the design of multi-functional wireless and satellite communication systems. These tunable components provide compactness and cost reduction to the RF front-ends since they combine multiple bands in one unit to save space and material. There have been techniques reported such as radio frequency microelectromechanical switches (RF-MEMS) [46,47], ferromagnetic materials [48], or semiconductor varactors [49–51] to design tunable filters. Among those techniques, ferromagnetic material and RF-MEMS-based tunable filters are expensive and the fabrication process can be challenging. Semiconductor varactor loaded microstrip tunable filters, on the other hand,
are compact, low-cost, and easy to integrate. All these advantages make varactor loaded filters very attractive to be used in multi-band wireless front-ends.

To further miniaturize the filter design, we choose to use planar split ring resonators (SRRs). SRRs are known for their compactness since they resonate at a dimension much smaller than a wavelength [11]. They are high-Q resonators and can be excited by a time-varying electric or magnetic field. There are a number of studies on varactor loaded split ring resonators. The transmission through capacitor loaded SRR has been investigated for various capacitance values [52]. A tunable bandpass tunable filter based on varactor loaded SRRs (VLSRRs) is presented by Gil [53] with a fairly high insertion loss. The basic tunable bandpass filter module that we use to design the third-order filter in this work is similar to VLSRR, except in our module the separation between the split rings of the SRR is uniform. The purpose of this study is to present a low-loss tunable bandpass filter using varactor loaded SRRs.

4.2 Design of the Tunable Microstrip Filter

The proposed tunable filter is designed by placing a tuning element between the two concentric rings of a SRR. The tuning element is a reverse-biased varactor diode and its capacitance can be tuned by changing the DC voltage applied to its pins. The resonant frequency of the SRR depends on the equivalent capacitance value between the split rings at the resonance.

4.2.1 Basic Tunable Filter Module

A tunable filter module with resonant frequency located in the vicinity of 2.7 GHz is designed. The dimensions of the basic tunable block are shown in Fig. 4.1. The spacing between concentric rings of SRR is set to 1.5 mm to leave enough space for a tuning diode. The width of the gap $w_4$ is set narrow enough to get an insertion loss less than 2 dB at the passband. Rogers RO4003 high frequency laminate (relative permittivity=3.38, thickness=1.524 mm) is used as the substrate. The characteristic impedance of the input line is set to 50 Ohm. The commercially available tuning diode Infenion BB833 is chosen
as the varactor. The varactor diode junction capacitance can be tuned between 9.8 pF and 0.8 pF by changing its biasing voltage. The excitation and coupling between the SRRs are through the magnetic field. The tuning diode is placed between two concentric split rings of each SRR so that it has minimum effect on the excitation and coupling between the SRRs.

The simulation results (performed with Agilent’s Momentum) for the scattering parameters of the basic tunable bandpass filter module for various varactor capacitance values are illustrated in Fig. 4.2. In the simulation, an ideal varactor that does not permit current to leak into the inner ring of the SRR is placed between concentric rings of the SRR. When the varactor capacitance is varied from 9.8 pF to 0.8 pF, the resonant frequency is tuned from 2.73 GHz to 2.61 GHz. The simulated insertion loss is low and constant through the tuning range.

The measured scattering parameters of the basic module for various biasing voltages are illustrated in Fig. 4.3. The measurements are performed on a HP8510 C vector analyzer. As the biasing voltage of the diode is varied from 0 V to 25 V, the passband frequency of the basic tunable filter module moves from 2.62 GHz to 2.75 GHz. The measured insertion loss is 3.1 dB at the highest frequency and 9.8 dB at the lowest frequency. The simulation
Fig. 4.2: Simulated $S_{11}$ and $S_{21}$ parameters of the basic tunable filter module.

results (Fig. 4.2) agree reasonably well with the measurements (Fig. 4.3). The reason for the difference of insertion losses in the lower frequencies (Fig. 4.3) is as follows. As the resonant frequency is decreased by lowering the varactor biasing voltage, current starts to leak into the inner split ring from outer ring through the diode varactor, which increases the insertion loss of the tunable module.

Fig. 4.3: Measured $S_{11}$ and $S_{21}$ parameters of the basic tunable filter module.
4.2.2 Multi-Stage Tunable Bandpass Filter Implementation

The filter synthesis using coupled resonators approach presented in Hong and Lancaster [29] is applied to design the tunable filter. We used the generic coupled resonators scheme of third-order bandpass filter structure with transmission zeroes located at the upper side of the passband. The design is performed by calculating the coupled resonator-based bandpass filter parameters, i.e., coupling coefficients and external quality factor presented in Hong and Lancaster [29]. A third-order tunable bandpass filter using varactor loaded SRRs with a passband centered at 2.4 GHz is designed. The layout and dimensions of the filter are shown in Fig. 4.4.

The simulated and measured scattering parameter results of the third-order tunable bandpass filter are depicted in Fig. 4.5 and Fig. 4.6, respectively. The fabricated filter can be tuned from 2.11 GHz up to 2.34 GHz by changing the biasing voltage from 0 V to 25 V. The measured insertion loss at 2.11 GHz is 11.65 dB and 1.98 dB at 2.34 GHz. The insertion loss difference between the lowest and highest frequency is 9.67 dB. For the applications where such a difference is acceptable, the tuning range and bandwidth of the filter are shown in Fig. 4.4.

Fig. 4.4: Schematic of the third-order tunable bandpass filter: \( d_1 = d_4 = 1 \text{ mm, } d_2 = 0.85 \text{ mm, } d_3 = 2.05 \text{ mm, } w_1 = 1.5 \text{ mm, } w_2 = 1 \text{ mm, } w_3 = 12 \text{ mm, } w_4 = 2.9 \text{ mm.} \) All the resonators are the same size.
filter are calculated to be 10.8% and 2%, respectively, using the measured $S_{21}$ data of the fabricated filter shown in Fig. 4.6. The insertion loss of the filter is measured as 6.18 dB at 2.2 GHz which is 4.2 dB lower than insertion loss at 2.34 GHz. In applications where an insertion loss less than 5 dB is needed through the tuning range, tuning range of the fabricated third-order tunable bandpass filter is 6.2%. Figure 4.7 shows the fabricated third-order tunable bandpass filter printed on a Rogers RO4003 substrate. Despite the high insertion loss in the lower frequency band, the tuning range of the varactor loaded SRR filter is fairly good and is five times of the filter bandwidth.
4.3 Conclusion

A compact planar tunable microstrip bandpass filter using varactor loaded SRRs has been implemented. A reverse-biased diode with voltage dependent variable nonlinear capacitance is used as the varactor. A prototype third-order tunable bandpass filter using three coupled resonator with a passband located at around 2.4 GHz is designed, fabricated, and tested. The dimension of the third-order fabricated tunable filter is 23.8 mm by 24.9 mm and is less than 1/3 of the resonant wavelength. Therefore, the miniaturization factor is about 3.5 compared to a third-order coupled line filter.
Chapter 5
Dual- and Triple-Band Wilkinson Power Dividers Based on Composite Right- and Left-Handed Transmission Lines

Abstract

This paper presents dual- and triple-band equal-split Wilkinson power dividers based on composite right- and left-handed transmission lines (CRLH TLs). Good isolation between output ports and impedance matching at input ports are achieved simultaneously at all passband frequencies for both dual- and triple-band Wilkinson power dividers. The theoretical closed-form design equations are derived, and the design allows a wide range of flexibility for the allocations of the passbands. To verify the proposed design, two prototypes are fabricated and measurements show good agreement with simulated data. The first design is a triple-band Wilkinson power divider with passbands centered at 0.8 GHz, 1.3 GHz, 1.85 GHz, and the second one is a dual-band Wilkinson power divider with passbands centered at 0.7 GHz, 1.5 GHz. The triple-band divider has a length of 0.66 wavelength in the substrate and is more compact comparing with traditional Wilkinson power dividers. The dual-band power divider is designed to have a wide fractional bandwidth with more than 20% at the lower passband and 41% at the upper passband. Measurements also show a good insertion loss at each input port, which is less than 3.6 dB at the center of each passband.

5.1 Introduction

Wilkinson power dividers [54] are important components of microwave and antenna systems. They are commonly used for antenna array feeds [55, 56], power amplifiers [57, 58], and mixers [59] because of their high isolation between output ports. While very popular, Wilkinson power dividers based on conventional transmission lines are bulky and
create challenges for the miniaturization of microwave systems. Another limitation of a conventional Wilkinson power divider is that when operating at two bands, the second band is usually restricted to the odd harmonics of the fundamental frequency. With the increasing demand for compact multi-band Wilkinson power dividers, many researchers have explored methods for miniaturization and for supporting multi-band operation.

A numerically near-exact solution for a dual-band Wilkinson power divider operating at fundamental frequency ($f_0$) and its first even harmonic ($2f_0$) has been presented [60]. The work showed the feasibility of moving the second band from $3f_0$ to $2f_0$; however, two frequency bands are closely related. Analytically exact solutions for equal-split dual-band Wilkinson power dividers operating at two arbitrary frequencies has been demonstrated, and lumped inductors and capacitors were used in parallel with isolation resistors to achieve ideal Wilkinson power divider responses [61,62]. Dual-band Wilkinson power dividers based on input stubs and cascaded transmission lines were presented by Cheng and Wu [63–65]. Recently, a triple-band Wilkinson power divider using three-section transmission line transformers was reported [66]. An unequal-split Wilkinson power divider with more than two output ports was reported [67], and a design of dual-band unequal-split Wilkinson power divider was presented by Wu [68]. All of those designs were based on conventional microstrip transmission lines, and the designs exhibit degrees of limitation such as relatively narrow band [61,66] and non-exact design equations [66].

A composite right- and left-handed transmission line (CRLH TL) is a transmission line approach of metamaterials [10], and has gained great interest in the design of novel microwave devices because of its broad bandwidth, low loss, and an unique dispersion characteristic [16,69–73]. Unlike the conventional right-handed transmission lines, the phase slope and zero phase frequency of CRLH TLs can be flexibly designed, so the frequency of the third harmonic of a CRLH TL is not necessary three times of the fundamental frequency ($f_0$) [21], suggesting a novel approach for multi-band circuitry. The flexibility of designing the zero phase frequency enables a CRLH TL to have two arbitrary phase values at two arbitrary frequencies ($f_1, f_2$) as long as $f_2 < 3f_1$. A broadband power divider with
more than two output ports was designed using zero-degree metamaterial lines, which is another transmission line approach of metamaterials [74]. In this paper, we present a novel dual- and tri-band equal-split Wilkinson power divider using CRLH TLs with exact design equations. The design equations for the isolation of the output ports are also presented. A triple- and a dual-band Wilkinson power divider were designed, fabricated, and tested for demonstration. The proposed dual-band power divider has a wider fractional bandwidth (FBW) than previously reported designs, and the size of the triple-band power divider is reduced to 3/4 of the triple-band Wilkinson power divider by Chongcheawchamnan [66].

5.2 Theory and Design Equations

The schematic of the proposed power divider is presented in Fig. 5.1. The divider consists of two pieces of CRLH TLs at each branch and two lumped resistors ($R_1$ and $R'_1$) for isolation between the output ports 2 and 3. The same topology is used for both dual- and triple-band Wilkinson power dividers. The conventional design methodology of Wilkinson power dividers explained by Pozar is used to calculate the branch input impedances ($Z_{in1}$, $Z_{in2}$) and output impedances ($R_2$, $R_3$) [75]. The two output ports of the divider are driven by symmetric (even-mode) and asymmetric (odd-mode) voltage sources [75]. The electrical lengths and characteristic impedances of the CRLH TLs are obtained from even-mode analysis, and the isolation resistors are determined from odd-mode analysis of the power divider [75]. In order to obtain equal power division, the circuit needs to be symmetric (i.e., $Z_{11}=Z_{21}$, $Z_{12}=Z_{22}$, $\theta_{11}=\theta_{21}$, and $\theta_{12}=\theta_{22}$ in Fig. 5.1). Each CRLH TL section has to be designed such that each branch fully transforms the input impedance to the output impedances simultaneously at the center frequencies of all passbands.

5.2.1 CRLH TLs

The schematic of the CRLH TL unit cell used for the Wilkinson power divider design is presented in Fig. 5.2. A left-handed transmission line (LH TL) [76] is 1-D implementation of left-handed media where the Poynting vector and the phase velocity are antiparallel, therefore, resulting in negative permeability and permittivity [10]. A typical CRLH TL
Fig. 5.1: Schematic of the proposed Wilkinson power divider.

Fig. 5.2: The schematic of the CRLH TL unit cell \((N=1)\) used in the proposed power divider.
can be assembled from \( N \) unit cells that are composed of left-handed (LH) and right-handed (RH) sections [73]. The left-handed section is based on a shunt inductor and a series capacitor and a right-handed section implemented with a conventional microstrip transmission line, a series inductor, and a shunt capacitor (Fig. 5.2). The inherent cut-off frequencies of the left- and the right-handed sections of the CRLH TL unit cell can be computed from

\[
f_{c}^{RH} = \frac{1}{\pi k \sqrt{L_{R,LE}C_{R,LE}}}, \quad f_{c}^{LH} = \frac{1}{4\pi k \sqrt{L_{L}C_{L}}}. \tag{5.1}
\]

When the series and the shunt resonances of a CRLH TL unit cells are equal to each other, the CRLH TL is said to be balanced. The phase response (\( \theta \)) and the characteristic impedance (\( Z \)) of a balanced CRLH TL are

\[
\theta = \theta_{R} + \theta_{L} = \theta_{R,TL} + \theta_{R,LE} + \theta_{L} = -N2\pi f^{\sqrt{L_{R}C_{R}}} + \frac{N}{2\pi f^{\sqrt{L_{L}C_{L}}}} \tag{5.2}
\]

\[
\text{and}
\]

\[
Z = \sqrt{\frac{L_{R,TL}}{C_{R,TL}}} = \sqrt{\frac{L_{R,LE}}{C_{R,LE}}} = \sqrt{\frac{L_{L}}{C_{L}}}, \tag{5.3}
\]

where \( \theta_{R} \) and \( \theta_{L} \) are the phase responses of right- and left-handed sections, respectively. The indexes \( R_{T,L} \) and \( R_{L,LE} \) denote transmission-line-based right-handed and lumped-element-based right-handed. The coefficient \( k \) represents the ratio of the right-handed section in the CRLH TL. For example, \( k=0 \) means a pure transmission line implementation, and \( k=1 \) means a pure lumped element implementation.

It should be noted that \( f_{c}^{LH} \) needs to be smaller than the lowest passband frequency (\( f_{l} \)), and \( f_{c}^{RH} \) needs to be larger than the highest passband frequency (\( f_{h} \)). If \( f_{c}^{LH} \geq f_{l} \), then a larger number of unit cells (\( N \)) needs to be chosen, and if \( f_{c}^{RH} \leq f_{h} \), then a smaller \( k \) needs to be chosen. If \( N \) is increased, even though the ratio \( L_{L}/C_{L} \) and the characteristic impedance stay constant, then the values of \( L_{L} \) and \( C_{L} \) will decrease, therefore, resulting in a decrease in the cut-off frequency of the left-handed section.
5.2.2 Even-Mode Analysis

For the even-mode excitation, two in-phase signals with the same amplitude are applied to the ports 2 and 3 (Fig. 5.3) [75]. Since there is no current flowing through the plane of symmetry, the power divider can be intersected as shown in Fig. 5.3. Lumped components $R_1$ and $R'_1$ can consequently be omitted and the impedance at input is doubled, i.e., $Z_{in1}=Z_{in2}=2Z_0$. The input impedance must be simultaneously transformed to the output impedance at all passband frequencies.

We begin our analysis for triple-band operation, and then the design equations are simplified for dual-band operation. In order to determine the design equation of the divider, Monzons two-section transformer theorem is applied to the circuit in Fig. 5.3 [77]. The impedances seen from the central plane toward the input and output are calculated from

$$Z'_{in1, left} = Z_{11} \frac{Z_{in1} - j Z_{11} \tan \theta_{11}}{Z_{11} - j Z_{in1} \tan \theta_{11}}, \quad (5.4)$$

and

$$Z'_{in1, right} = Z_{12} \frac{R_2 + j Z_{12} \tan \theta_{12}}{Z_{12} + j R_2 \tan \theta_{12}}, \quad (5.5)$$

By equating $Z'_{in1, right}$ to $Z'_{in1, left}$ and solving (5.4) and (5.5) by equating the real and imaginary parts, we obtain the following two equations:

$$\tan[\theta_{11}(f)] \tan[\theta_{12}(f)] = \frac{Z_{11} Z_{12}(R_2 - Z_{in1})}{(Z_{11}^2 R_2 - Z_{12}^2 Z_{in1})}, \quad (5.6)$$

and

$$\frac{\tan[\theta_{11}(f)]}{\tan[\theta_{12}(f)]} = \frac{Z_{11}(Z_{12}^2 - R_2 Z_{in1})}{Z_{12}(R_2 Z_{in1} - Z_{11}^2)}, \quad (5.7)$$

After substituting $f$ with $f_1$, $f_2$, and $f_3$ in (5.6) and (5.7), one can determine that (5.8) and (5.9) must be fulfilled in order to match input impedance simultaneously at all passbands.
\[
\tan[\theta_{11}(f_1)] = \pm \tan[\theta_{11}(f_2)] = \pm \tan[\theta_{11}(f_3)]
\] (5.8)

\[
\tan[\theta_{12}(f_1)] = \pm \tan[\theta_{12}(f_2)] = \pm \tan[\theta_{12}(f_3)]
\] (5.9)

From (5.8) and (5.9), we have the following equations:

\[
\theta_{11}(f_1) \pm \theta_{11}(f_2) = -l\pi,
\] (5.10)

\[
\theta_{11}(f_2) \pm \theta_{11}(f_3) = -m\pi,
\] (5.11)

\[
\theta_{11}(f_1) \pm \theta_{11}(f_3) = -n\pi,
\] (5.12)

\[
\theta_{12}(f_1) \pm \theta_{12}(f_2) = -p\pi,
\] (5.13)

\[
\theta_{12}(f_2) \pm \theta_{12}(f_3) = -r\pi,
\] (5.14)

\[
\theta_{12}(f_1) \pm \theta_{12}(f_3) = -s\pi,
\] (5.15)

where \(l, m, n, p, r,\) and \(s\) are arbitrary integers. If we define the required phase response of first CRLH TL section as

\[
\theta_{11}(f_1) = -\frac{\pi}{2} + \phi,
\] (5.16)

where \(\phi\) is a variable, then the phase response of \(\theta_{11}\) at \(f_3\) becomes

\[
\theta_{11}(f_3) = \theta_{11}(f_1) - \pi = -\frac{3\pi}{2} + \phi,
\] (5.17)
since the third harmonic of a CRLH TL is $180^0$ out of phase with its first harmonic.

If the plus operation in (5.10) is picked and $l$ is assigned as 1, then $\theta_{11}(f_2)$ becomes

$$\theta_{11}(f_2) = -\frac{\pi}{2} - \phi. \quad (5.18)$$

After substituting $\theta_{11}(f_3)$ and $\theta_{11}(f_2)$ into (5.11) and picking the plus operation in (5.11), one has

$$m = \left(-3\pi/2 + \phi - \pi/2 - \phi\right)/\pi = 2. \quad (5.19)$$

After substituting $\theta_{11}(f_1)$ and $\theta_{11}(f_3)$ into (5.12) and picking the minus operation, $n$ can be calculated from

$$n = \left(-\pi/2 + \phi + 3\pi/2 - \phi\right)/\pi = -1. \quad (5.20)$$

If we do the same process to (5.13)-(5.15) as we did for (5.10)-(5.12) and choose $p=l$, $r=m$, and $s=n$, then we have

$$\theta_{11}(f_1) = \theta_{12}(f_1), \quad (5.21)$$

$$\theta_{11}(f_2) = \theta_{12}(f_2), \quad (5.22)$$

and

$$\theta_{11}(f_3) = \theta_{12}(f_3), \quad (5.23)$$

for the phase response of CRLH TL sections.

If we choose

$$\sqrt{L_{R11}C_{R11}} = \sqrt{L_{R12}C_{R12}} \quad \text{and} \quad \sqrt{L_{L11}C_{L11}} = \sqrt{L_{L12}C_{L12}},$$

by using (5.2), the design equations (5.16)-(5.18) can be rewritten as
\[-N2\pi f_1 \sqrt{L_{R11}C_{R11}} + \frac{N}{2\pi f_1 \sqrt{L_{L11}C_{L11}}} - \phi = -\frac{\pi}{2}, \quad (5.24)\]

\[-N2\pi f_2 \sqrt{L_{R11}C_{R11}} + \frac{N}{2\pi f_2 \sqrt{L_{L11}C_{L11}}} + \phi = -\frac{\pi}{2}, \quad (5.25)\]

and

\[-N2\pi f_3 \sqrt{L_{R11}C_{R11}} + \frac{N}{2\pi f_3 \sqrt{L_{L11}C_{L11}}} - \phi = -\frac{3\pi}{2}. \quad (5.26)\]

After some algebra, we obtain from (5.24)-(5.26) the following:

\[\sqrt{L_{R11}C_{R11}} = \frac{f_3/(f_3 - f_1) + f_2/(f_1 + f_2)}{2N(f_2 + f_3)}, \quad (5.27)\]

\[\sqrt{L_{L11}C_{L11}} = \frac{N(f_1 + f_2)(f_2 + f_3)(f_3 - f_1)}{2\pi^2 f_1 f_2 f_3 (2f_1 + f_2 - f_3)}, \quad (5.28)\]

and

\[\phi = \frac{\pi}{2} \frac{f_2 f_3 (3f_2 - f_3) + f_1 f_2 (f_1 - f_2) + f_1 f_3 (f_3 - 3f_1)}{(f_3 - f_1)(f_1 + f_2)(f_2 + f_3)}. \quad (5.29)\]

The characteristic impedances of each section can be computed according to (Monzon) [77] as

\[Z_{11} = \sqrt{\frac{Z_{in1}}{2\alpha} (R_2 - Z_{in1}) + \sqrt{\left[\frac{Z_{in1}}{2\alpha} (R_2 - Z_{in1})\right]^2 + Z_{in1}^3 R_2}}, \quad (5.30)\]

\[Z_{12} = \frac{Z_{in1} R_2}{Z_{11}}. \quad (5.31)\]

For dual-band operation, the first and second passband superpose (i.e., \(f_1 = f_2\)), so the phase responses (5.16) and (5.18) become identical to each other (i.e., \(\phi = 0\)). Each CRLH TL section becomes a quarter-wave transformer, so the characteristic impedances of
CRLH TLs can be determined by using the quarter-wave transformer impedance equations to have [75]:

\[ Z_{11} = Z_{in1}^{3/4}Z_0^{1/4}, \]  

(5.32)

and

\[ Z_{12} = Z_0^{3/4}Z_{in1}^{1/4}. \]  

(5.33)

The equations (5.27) and (5.28) are reduced to

\[ \sqrt{L_{R11}C_{R11}} = \frac{2f_2/(f_2 - f_1) + 1}{4N(f_1 + f_2)}, \]  

(5.34)

and

\[ \sqrt{L_{L11}C_{L11}} = \frac{N(f_1 + f_2)(f_2 - f_1)}{\pi^2f_1f_2(3f_1 - f_2)}, \]  

(5.35)

where \( f_1 \) and \( f_2 \) are the center frequencies of the first and second bands, respectively [69].

The physical length of the microstrip transmission lines (\( l_{TL} \)) for each CRLH TL section can be calculated using standard microstrip transmission line formulas for known parameters \( \sqrt{L_{R11}C_{R11}}, k, f_1 \) and using (5.2) [75].

### 5.2.3 Odd-Mode Analysis

For odd-mode excitation, two signals with same magnitude and \( 180^0 \) phase difference are applied to the ports 2 and 3. In this type of excitation, analytically, there is a virtual ground along the plane of symmetry, and the input impedance becomes zero [75]. Accordingly, the proposed power divider can be bisected as shown in Fig. 5.4.
Fig. 5.3: The equivalent circuit of the proposed power divider for even-mode analysis.

Fig. 5.4: The equivalent circuit of the proposed power divider for odd-mode analysis.
To perfectly isolate the output ports, the impedance seen from the output ports 2 and 3 toward the input must be equal to output impedances; i.e., \( R_2 = R_3 = Z_0 \) has to be held simultaneously at all passbands [75]. Since the circuit is symmetric, one can only analyze one output port and obtain the information for other ports easily from the analysis. So the calculation for only port 2 is presented in the following. The impedance seen from the output of the equivalent odd-mode circuit shown in Fig. 5.4 is

\[
Z_{\text{odd}1} = Z_{12} \frac{Z'_{\text{odd}1} + jZ_{12}\tan\theta_{12}}{Z_{12} + jZ'_{\text{odd}1}\tan\theta_{12}} / r_1', \quad (5.36)
\]

and the impedance seen from the central plane toward the input is

\[
Z'_{\text{odd}1} = Z_{11} \frac{0 + jZ_{11}\tan\theta_{11}}{Z_{11} + j0\tan\theta_{11}} / r_1 = jZ_{11}\tan\theta_{11} / r_1. \quad (5.37)
\]

By equating \( Z_{\text{odd}1} \) to \( R_2 \), and solving (5.36), (5.37) for \( r_1' \) and \( r_1 \), we obtain that (5.38) and (5.39) have to be satisfied for the perfect isolation of the outputs.

\[
r_1' r_1 (Z_{11} + Z_{12}) = R_2 (r_1 Z_{11} + r_1 Z_{12} - r_1' Z_{11}) \quad (5.38)
\]

\[
R_2 r_1' r_1 (Z_{12} - Z_{11}\tan\theta_{11}\tan\theta_{12}) = Z_{11}Z_{12}^2 (R_2 - r_1')\tan\theta_{11}\tan\theta_{12} \quad (5.39)
\]

From (5.16)-(5.18) and (5.21)-(5.23), one can determine that the relation holds for all passbands, and it implies that (5.38) and (5.39) hold for all passband frequencies for appropriate \( r_1 \) and \( r_1' \). In order to determine \( r_1 \) and \( r_1' \), we introduce the following terms:

\[
a = R_2^2 Z_{11} (Z_{12} - Z_{11}(\tan\theta_{11})^2) + (Z_{11} + Z_{12})Z_{11}Z_{12}^2(\tan\theta_{11})^2, \quad (5.40)
\]

\[
b = -2R_2 (Z_{11} + Z_{12})Z_{11}Z_{12}^2(\tan\theta_{11})^2, \quad (5.41)
\]

and
\[ c = R_2^2(Z_{11} + Z_{12})Z_{11}^2Z_{12}(\tan\theta_{11})^2. \]  
\hspace{2cm} (5.42)

With (5.40)-(5.42), from (5.38) and (5.39), one can reach a parabolic equation for \( r'_1 \):

\[ a(r'_1)^2 + b(r'_1) + c = 0. \]  
\hspace{2cm} (5.43)

The two solutions for (5.43) are

\[ r'_1 = (-b \pm \sqrt{\Delta})/2a, \]
\hspace{2cm} (5.44)

where \( \Delta = b^2 - 4ac \).

In dual-band operation where \( \phi = 0 \) and \( \tan\theta_{11} = \infty \), l'Hopital's rule is applied on (5.44) with respect to \( \tan\theta_{11} \) to yield (5.45).

\[ r'_1 = R_2 + \frac{R_2^2Z_{11}^2}{(Z_{11} + Z_{12})Z_{11}Z_{12} - R_2^2Z_{11}^2} \pm \]
\[ \sqrt{\frac{R_2^2 + (Z_{11} + Z_{12})^2Z_{11}^2Z_{12}^2 - (Z_{11} + Z_{12})Z_{11}Z_{12}^2 - R_2^2Z_{11}Z_{12}^2(Z_{11} + Z_{12})R_2^2Z_{11}Z_{12}^2}{(Z_{11} + Z_{12})^2Z_{11}^2Z_{12}^2 + R_2^2Z_{11}^2Z_{12}^2 - 2(Z_{11} + Z_{12})Z_{11}Z_{12}^2R_2^2Z_{11}^2}}. \]  
\hspace{2cm} (5.45)

Since \( Z_{11}, Z_{12}, \) and \( R_2 \) are known from the previous calculations, one can compute \( r'_1 \) from (5.45), then find \( r_1 \) from (5.38) or (5.39). It should be noted that \( r_1 \) and \( r'_1 \) have to be positive numbers, and therefore, the plus or minus sign in (5.39) has to be chosen accordingly. It can be seen that \( \Delta \) can become negative and yield complex roots for \( r'_1 \). For those cases, (5.44) cannot be used to calculate the isolation resistors. Instead, the resistors \( R_1 \) and \( R'_1 \) can be designed such that the output ports are matched at all passbands as described as follows.

First, we set \( r_1 = \infty \), and calculate \( Z'_{\text{odd}} \) from (5.37). The value for \( Z_{\text{odd}} \) is then computed from (5.36). Because \( R_2 = Z_0 = Z_{\text{odd}} \), we have

\[ R_2 = \frac{jr'_1Z_{12}\tan\theta_{11}(Z_{11} + Z_{12})}{j(Z_{12}\tan\theta_{11}(Z_{11} + Z_{12})) + r'_1(Z_{12} - Z_{11}(\tan\theta_{11})^2)}. \]  
\hspace{2cm} (5.46)
For frequencies when $\theta_{11}$ is around $45^0$, then we have

$$r'_1 = (Z_{12} - Z_{11}(\tan\theta_{11})^2) << Z_{12}\tan\theta_{11}(Z_{11} + Z_{12}), \quad (5.47)$$

and (5.46) can be reduced to

$$r'_1 \cong R_2. \quad (5.48)$$

Finally, $R_1$ and $R'_1$ can be calculated from

$$R_1 = 2r_1 \text{ and } R'_1 = 2r'_1, \quad (5.49)$$

since the circuit is symmetric. For passband frequencies when $\Delta$ is negative and the condition in (5.47) cannot be fulfilled, one cannot achieve a good isolation between output ports.

### 5.3 Implementation and Experimental Results

To verify the design methodology, a triple-band and a dual-band equal-split power divider were designed, fabricated, and tested. The design procedure is summarized as follows.

- Decide $N$ and $k$ according to sec. 5.2.1 for desired passbands. The passbands needs to be between $f_c^{LH}$ and $f_c^{RH}$. Otherwise, one needs to adjust the CRLH TL design by increasing $N$ or decreasing $k$ until the passbands fall between $f_c^{LH}$ and $f_c^{RH}$.

- Calculate $\sqrt{L_{R11}C_{R11}}$, $\sqrt{L_{R12}C_{R12}}$, and using (5.27)-(5.28) for triple-band or (5.34)-(5.35) for dual-band.

- Find the characteristic impedance and the electrical length of the sections in each branch of divider according to (5.30)-(5.31) for triple-band, or (5.32)-(5.33) for dual-band.
• Calculate the values of the lumped inductors and capacitances for the chosen $k$ and $N$.

• Solve (5.2) for $\theta_{R,TL11}$ and $\theta_{R,TL12}$ by substituting $\sqrt{L_{R11}C_{R11}}$, $\sqrt{L_{R12}C_{R12}}$, $k$, $N$, and $f_1$.

• Determine the electrical length and the line widths of the right-handed transmission line section of CRLH TLs using the standard microstrip transmission line formula.

• Calculate $r'_1$ and find isolation resistors $R_1$ and $R'_1$ according to sec. 5.2.3.

The design and simulations were performed with Agilent’s ADS. A LPKF 93S milling machine is used to fabricate the Wilkinson power divider boards. The RO4003 laminates ($\varepsilon_r=3.38$, $h=1.524$ mm, and $\tan\delta=0.0021$) were used as substrates, and $Z_0$ is set to 50 $\Omega$ for both designs. The lumped elements (capacitances and inductances) are from the Murata Manufacturing Co. Ltd. The capacitances are selected from LQW series, and the inductors are from GJM series. The self-resonant frequencies of the selected lumped elements are at least two times larger than the highest passband. Measurements of the scattering parameters were performed using an Agilent/HP 8510 vector network analyzer.

5.3.1 Triple-Band Equal-Split Wilkinson Power Divider

The center frequencies of the three passbands of the triple-band equal-split power divider are designed to be $f_1=$0.8 GHz, $f_2=$1.3 GHz, and $f_3=$1.85 GHz. $N$ and $k$ are decided to be 3 and 0.6, respectively, according to sec. 5.2.1. The design parameters of the triple-band Wilkinson power divider are presented in Table 5.1. The designed values for lumped elements and values that are actually available from the manufacturer are also listed in Table 5.1. The characteristic impedances and electrical lengths of the CRLH TL sections in each branch of the divider are calculated from (5.30)-(5.31). The values of the isolation resistors are obtained from (5.48) and (5.49). The values of $L_{L11}$, $C_{L11}$, $L_{L12}$, $C_{L12}$, $kL_{R11}$, $kC_{R11}$, $kL_{R12}$, and $kC_{R12}$ are computed from (5.27)-(5.29) and (5.3). Finally,
Table 5.1: Design parameters of the triple-band equal-split Wilkinson power divider $f_1 = 0.8$ GHz, $f_2 = 1.3$ GHz, and $f_3 = 1.85$ GHz ($N=3$, $k=0.6$).

<table>
<thead>
<tr>
<th>$Z_{11}, Z_{21}$</th>
<th>$Z_{12}, Z_{22}$</th>
<th>$R_1, R'_1$</th>
<th>$\theta_{11}, \theta_{12}$</th>
<th>$Z_{in1}, Z_{in2}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>69.3 Ω</td>
<td>72.1 Ω</td>
<td>$\infty, 100 \Omega$</td>
<td>0.24 $\pi$ at $f_1$</td>
<td>100 Ω</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Lumped Elements</th>
<th>Calculated</th>
<th>Used</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{L11}$</td>
<td>36.23 nH</td>
<td>3x12 nH</td>
</tr>
<tr>
<td>$C_{L11}, 2C_{L11}$</td>
<td>7.54 pF, 15.08 pF</td>
<td>2x3.8 pF, 4x3.8 pF</td>
</tr>
<tr>
<td>$L_{L12}$</td>
<td>37.68 nH</td>
<td>3x13 nH</td>
</tr>
<tr>
<td>$C_{L12}, 2C_{L12}$</td>
<td>7.25 pF, 14.5 pF</td>
<td>2x3.6 pF, 4x3.6 pF</td>
</tr>
<tr>
<td>$kL_{R11}, kL_{R11}/2$</td>
<td>5.24 nH, 2.62 nH</td>
<td>5.1 nH, 2.7 nH</td>
</tr>
<tr>
<td>$kC_{R11}$</td>
<td>1.09 pF</td>
<td>1.0 pF</td>
</tr>
<tr>
<td>$kL_{R12}, kL_{R12}/2$</td>
<td>5.44 nH, 2.72 nH</td>
<td>5.6 nH, 2.7 nH</td>
</tr>
<tr>
<td>$kC_{R12}$</td>
<td>1.05 pF</td>
<td>1.1 pF</td>
</tr>
</tbody>
</table>

$l_{TL11}, l_{TL12}, w_{TL11},$ and $w_{TL12}$ are calculated to be 14.1 mm, 14.2 mm, 1.96 mm, and 1.81 mm, respectively.

The fabricated Wilkinson power divider is shown in Fig. 5.5. The dimension of the divider is 7.6 cm by 1.9 cm, which corresponds to 0.66 $\lambda_1$ by 0.16 $\lambda_1$, where $\lambda_1$ is the guided wavelength at $f_1$ (0.8 GHz) in the RO4003 substrate. The simulated and measured group delay and insertion loss of the triple-band power divider are presented in Fig. 5.6. Figure 5.7 shows the simulated and measured output isolation and input reflection coefficients. The simulation and measurements for the insertion and input return losses, isolation between the outputs, and the FBW are listed in Table 5.2. The FBW for each passband is defined as the frequency range (both the input return loss and the isolation are less than -15 dB) divided by the center frequency of the passband. In the measurements, the passbands are shifted to 0.84 GHz, 1.28 GHz, and 1.68 GHz. These shifts are mainly due to the difference between the calculated and the used lumped elements, and the difficulty in fabricating the precise line-width for the CRLH TLs. The measured insertion loss is lower than 3.25 dB for all three passbands. The maximum variation of insertion loss (insertion amplitude imbalance) within the bandwidth is 0.12 dB, 0.18 dB, and 0.11 dB at passbands $f_1$, $f_2$, and $f_3$, respectively, indicating that the insertion loss within the passbands is reasonably constant and stable.
Fig. 5.5: Fabricated triple-band equal-split Wilkinson power divider with $f_1 = 0.8$ GHz, $f_2 = 1.3$ GHz, and $f_3 = 1.85$ GHz.

Fig. 5.6: Simulated and measured group delay and $S_{21}$ (insertion loss) parameters of the triple-band equal-split Wilkinson power divider.
Within the passbands $f_1$, $f_2$, and $f_3$, the maximum measured input return losses are 49.8 dB, 41.4 dB, and 28.9 dB, respectively, and the maximum measured output isolations are 23.3 dB, 25.1 dB, and 29 dB, indicating that good matching and isolation are achieved at the three passbands. The simulated and measured output return losses (ORLs) of the divider are presented in Fig. 4.8 and show that a good output port matching is achieved at all passbands. It is seen that the agreement between simulations and measurements are good at each passband. The measured and simulated group delays in the passbands are constant and show good agreement. The measured input return loss, output isolation, and output return loss are slightly different from simulations, and this can be due to the difference between the selected and the available lumped elements, parasitic effects of soldering, and the precision in the fabrication.

### 5.3.2 Dual-Band Equal-Split Wilkinson Power Divider

The dual-band equal-split power divider is designed for 0.7 GHz and 1.5 GHz. The value for $N$ and $k$ are 3 and 0.5. The design parameters are presented in Table 5.3. The values for $l_{TL11}$, $l_{TL12}$, $w_{TL11}$, and $w_{TL12}$ are found to be 25.5 mm, 25.6 mm, 1.31 mm, and
Fig. 5.8: Measured and simulated output return loss parameters of the triple-band equal-split Wilkinson power divider.

2.60 mm.

Figure 5.9 shows the implemented power divider with a dimension of 11.9 cm and 2.1 cm, which corresponds to 0.90 $\lambda_1$ by 0.16 $\lambda_1$, where $\lambda_1$ is the wavelength at $f_1$ in the RO4003 substrate. The simulated and measured insertion loss and group delay of the dual-band power divider are shown in Fig. 5.10. The measured and simulated group delay is constant within the passbands. Figure 5.11 shows the simulated and measured output isolation and input reflection coefficients. The rest of the performance parameters are summarized in Table 5.4. The definition of the FBW is the same as the one for the triple-band power

Table 5.2: Simulated and measured parameters of the triple-band equal-split Wilkinson power divider.

<table>
<thead>
<tr>
<th></th>
<th>Simulation</th>
<th>Measurement</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Frequency (GHz)</strong></td>
<td>$f_1$ 0.8</td>
<td>$f_1$ 0.84</td>
</tr>
<tr>
<td></td>
<td>$f_2$ 1.3</td>
<td>$f_2$ 1.28</td>
</tr>
<tr>
<td></td>
<td>$f_3$ 1.85</td>
<td>$f_3$ 1.68</td>
</tr>
<tr>
<td><strong>Insertion Loss (dB)</strong></td>
<td>3.03</td>
<td>3.21</td>
</tr>
<tr>
<td></td>
<td>3.04</td>
<td>3.12</td>
</tr>
<tr>
<td></td>
<td>3.11</td>
<td>3.23</td>
</tr>
<tr>
<td><strong>Maximum IRL (dB)</strong></td>
<td>34.9</td>
<td>49.8</td>
</tr>
<tr>
<td></td>
<td>37.4</td>
<td>41.4</td>
</tr>
<tr>
<td></td>
<td>25.2</td>
<td>28.9</td>
</tr>
<tr>
<td><strong>Maximum ORL (dB)</strong></td>
<td>36.4</td>
<td>35.0</td>
</tr>
<tr>
<td></td>
<td>36.1</td>
<td>26.8</td>
</tr>
<tr>
<td></td>
<td>25.1</td>
<td>26.4</td>
</tr>
<tr>
<td><strong>Output Isolation (dB)</strong></td>
<td>35.7</td>
<td>23.3</td>
</tr>
<tr>
<td></td>
<td>30.2</td>
<td>25.1</td>
</tr>
<tr>
<td></td>
<td>18.4</td>
<td>29.0</td>
</tr>
<tr>
<td><strong>FBW (%)</strong></td>
<td>17</td>
<td>14.7</td>
</tr>
<tr>
<td></td>
<td>11.5</td>
<td>12.5</td>
</tr>
<tr>
<td></td>
<td>7.8</td>
<td>6.5</td>
</tr>
</tbody>
</table>
divider in sec. 5.3.2. The agreement between simulations and measurements are good, and some slight discrepancies can be explained with the same reasons for the triple-band Wilkinson power divider.

The measured insertion amplitude imbalances are 0.16 dB and 0.39 dB for the lower and the upper passbands, respectively, which are satisfactory considering the wide FBW achieved by the design. Figure 5.12 shows the simulated and measured output port return losses, and it is seen that the simulated and measured output return losses are consistent. Good output isolation, output matching, and low input return loss is achieved in two passbands. It is shown Table 5.4 that the dual-band Wilkinson power divider has a wide fractional bandwidth and the measured FBW is larger than 20% at 0.7 GHz and 41% at 1.5 GHz.

5.4 Conclusion

Novel triple- and dual-band equal-split Wilkinson power dividers based on CRLH TLs are presented. Theoretical closed-form design equations are derived for arbitrary passbands. To validate the theoretical design equations, a dual- and a triple-band Wilkinson power dividers are designed, fabricated, and tested. Good isolation and input return loss are achieved for both dividers. The triple-band Wilkinson power divider is compact, and the

Table 5.3: Design parameters of the dual-band equal-split Wilkinson power divider $f_1 =0.7$ GHz and $f_2 =1.5$ GHz ($N=3$, $k=0.5$).

<table>
<thead>
<tr>
<th>$Z_{11},Z_{21}$</th>
<th>$Z_{12},Z_{22}$</th>
<th>$R_1,R_1'$</th>
<th>$\theta_{11},\theta_{12}$</th>
<th>$Z_{in1},Z_{in2}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>84.1 $\Omega$</td>
<td>59.5 $\Omega$</td>
<td>91$\Omega$, 280 $\Omega$</td>
<td>0.5 $\pi$ at $f_1$</td>
<td>100 $\Omega$</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Lumped Elements</th>
<th>Calculated</th>
<th>Used</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{L11}$</td>
<td>71.41 nH</td>
<td>4x18 nH</td>
</tr>
<tr>
<td>$C_{L11}, 2C_{L11}$</td>
<td>10.10 pF, 20.20 pF</td>
<td>2x5 pF, 4x5 pF</td>
</tr>
<tr>
<td>$L_{L12}$</td>
<td>50.49 nH</td>
<td>4x13 nH</td>
</tr>
<tr>
<td>$C_{L12}, 2C_{L12}$</td>
<td>14.28 pF, 28.56 pF</td>
<td>2x7.1 pF, 4x7.1 pF</td>
</tr>
<tr>
<td>$k L_{R11},k L_{R11}/2$</td>
<td>7.55 nH, 3.77 nH</td>
<td>7.5 nH, 3.9 nH</td>
</tr>
<tr>
<td>$k C_{R11}$</td>
<td>1.07 pF</td>
<td>1.1 pF</td>
</tr>
<tr>
<td>$k L_{R12},k L_{R12}/2$</td>
<td>5.34 nH, 2.67 nH</td>
<td>5.1 nH, 2.7 nH</td>
</tr>
<tr>
<td>$k C_{R12}$</td>
<td>1.56 pF</td>
<td>1.6 pF</td>
</tr>
</tbody>
</table>
Fig. 5.9: Fabricated dual-band equal-split Wilkinson power divider with $f_1 = 0.7 \text{ GHz}$ and $f_2 = 1.5 \text{ GHz}$.

Fig. 5.10: Simulated and measured group delay and $S_{21}$ (insertion loss) parameters of the dual-band equal-split Wilkinson power divider.
Fig. 5.11: Measured and simulated $S_{32}$ (output isolation) and $S_{11}$ (input reflection) parameters of the dual-band equal-split Wilkinson power divider. Black and blue double arrow lines represent measured and simulated bandwidths, respectively.

Fig. 5.12: Measured and simulated output return loss parameters of the dual-band equal-split Wilkinson power divider.
Table 5.4: Simulated and measured parameters of the dual-band equal-split Wilkinson power divider.

<table>
<thead>
<tr>
<th></th>
<th>Simulation</th>
<th>Measurement</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency (GHz)</td>
<td>$f_1$</td>
<td>$f_2$</td>
</tr>
<tr>
<td>Insertion Loss (dB)</td>
<td>3.07</td>
<td>3.11</td>
</tr>
<tr>
<td>Maximum IRL (dB)</td>
<td>45.5</td>
<td>46.4</td>
</tr>
<tr>
<td>Maximum ORL (dB)</td>
<td>46.7</td>
<td>24.4</td>
</tr>
<tr>
<td>Output Isolation (dB)</td>
<td>46.3</td>
<td>33.4</td>
</tr>
<tr>
<td>FBW (%)</td>
<td>49.4</td>
<td>26.3</td>
</tr>
</tbody>
</table>

dual-band Wilkinson power divider has wide fractional bandwidth. All measured results are in good agreement with the simulated results with slight variations that are explainable. The design concept can be extended to unequal-split multi-band Wilkinson power dividers. The proposed dividers are promising and effective components for multi-band microwave communication systems.
Chapter 6

Summary and Future Work

This chapter summarizes the dissertation and suggests possible future research that can be continued from the accomplished research.

6.1 Summary

This dissertation has demonstrated that metamaterial-inspired microwave circuits are very beneficial for bandwidth enhancement and miniaturization of multi-band microwave filters and power dividers.

It has been shown in Chapters 2 and 3 that split ring resonator (SRR), double split SRR (DS-SRR), and their complements (CSRR and DS-CSRR) are promising candidates for the miniaturization of dual-band bandpass filters. In Chapter 2, a novel way of integration of SRR and DS-CSRR to achieve dual-band response has been proposed and analyzed. In the following Chapter 3, we have presented a compact three-stage bandpass filter using SRR and DS-CSRR with the ability to design multi-stage filters with standard filter approximations. The characteristics of each passbands of the proposed dual-band filter can be designed individually. For demonstration, a prototype narrow-band dual-band Chebyshev bandpass filter with equal 0.4 dB ripples in both passbands has been designed and fabricated. Excellent out-of-band rejection and insertion loss performances have been achieved at both passbands. The measurements show close agreements with full-wave simulations and that proves the validity of the design methodology. Both SRR and DS-CSRR resonates at their fundamental frequencies. The proposed dual-band filter is a potential candidate for the miniaturization and cost-reduction of multi-band wireless microwave systems.

The main challenges of tunable bandpass filters are compactness and controlling the center frequency without any substantial change in the passband characteristics like frac-
tional bandwidth, out-of-band rejection ratio, and insertion loss. Chapter 4 presents a compact tunable coupled resonator bandpass filter based on varactor loaded split ring resonators with fairly high tuning range provided that its narrow bandwidth. The size of the varactor loaded microstrip tunable filters is 3.5 times smaller compared to quarter wavelength based-coupled filters.

Passive Wilkinson power dividers are widely used in antenna-array feedings, power amplifier input, and output networks. The size, isolation between the output ports, bandwidth, cost, input return, and input insertion losses of power dividers are very important performance parameters and need to be improved. In Chapter 5, we have presented equal-split 1:2 dual- and triple-band Wilkinson power dividers based on CRLH TLs. Due to nonlinear phase behavior of CRLH TLs, miniaturization and bandwidth enhancement in dual- and triple-band Wilkinson power dividers has been achieved. The design procedure and closed form analytical equations for input and output isolations for any arbitrary passband frequencies have been presented. The proposed divider has perfectly matched input and output ports has perfectly isolated outputs. Also the divider has low-cost, low-power, and substrate losses since it is built in microstrip with inexpensive lumped capacitors and inductors.

6.2 Future Work

SRRs, DS-CSRRs, and CRLH TLs have proven to be very useful for the design of miniaturized wideband, low-cost, low-loss, passive multi-band power dividers, dual-band, and tunable bandpass filters. However, there is still some room for research on improvement of passive and active microwave components. Some future research directions to follow the work presented herein can be summarized as follows.

In a tunable filter, semiconductor diodes have been used as varactors so that results in high insertion losses at passbands due to high current leakage between the SRR rings through diode junction. In order to solve this problem, the possible use of other tuning techniques on the proposed filter can be investigated.

Another area of research in the bandpass filters can be design of wideband multi-band
bandpass filters using CRLH TLs. The use of CRLH TLs in multi-band phase shifters, multipliers, and mixers can be studied. The presented Wilkinson power divider in Chapter 5 can be extended to unequal-split multi-output power dividers.

On the other hand, the design of metamaterial-inspired active microwave components is relatively unexplored compared to passive devices. CRLH TLs and SRRs can also be used to improve the performance of power amplifiers. So far numerous attempts have been made by researchers to miniature and to enhance power efficiency of the amplifiers. The conventional right-handed transmission-based input and output matching networks of power amplifiers are lack of matching the input and output besides at fundamental and its odd harmonic frequencies. That makes them not suitable for the design of multi-band power dividers with arbitrary frequency bands. CRLH TL-based input and output matching networks are beneficial to solve this problem since input and output matching can be achieved at two arbitrary frequencies. A dual-band class-E high efficiency power amplifier using CRLH TLs has been demonstrated [78]. CRLH TLs have been used in both input and output matching networks at dual frequencies. However, this design does not pay attention to the harmonic termination to improve power added efficiency. Another design of class-E dual-band power amplifiers using a CRLH TL with reduced size and power loss has been demonstrated [70]. The linearity and power efficiency are important performance parameters of multi-band power amplifiers and need to be improved. The CRLH TLs have been also used in output tuner of class-F type power amplifiers to cancel higher harmonics which mainly constitute the output nonlinearity [18]. An improved efficiency design of class-F power amplifier has been presented by reducing the parasitic effects of packaged power transistor by series capacitor and shunt inductor based compensation circuits [79,80]. There exist other various techniques to improve performance parameters such as Doherty and distributed amplifiers. A Doherty amplifier where main amplifier is set to class A and the peak amplifier is set to class C has been presented [81,82]. Compared to Doherty amplifiers designed using conventional right-handed transmission lines, significant enhancement in third-order output intermodulation products has been achieved at both designs. On the other hand, distributed
power amplifiers are commonly used in wireless systems when high output power capability and wide bandwidth are needed. Multiple power transistors are connected in parallel and the power is divided by each transistor. CRLH TLs have been used in distributed power amplifiers for the first time by Mata-Contreras [83,84] where right-handed transmission lines in gate and drain lines are replaced by CRLH TLs. Also, a dual-band distributed power amplifier is implemented [85] where gate RH TLs are replaced by CRLH TLs. In both designs of distributed power amplifiers, an improvement in linearity is presented. Another technique that has been used for providing high linearity is Chireix outphasing. CRLH TLs have been used in Chireix outphasing power amplifiers to suppress the second and third harmonics for enhancing the linearity [86]. The CRLH TLs can also been used in the design of triple-band class-F power amplifiers by replacing input output networks with two-section cascaded CRLH TLs.
References


Appendix
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Licensed content author Genc Alper, Baktur Reyhan
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<thead>
<tr>
<th>License Number</th>
<th>2371640884380</th>
</tr>
</thead>
<tbody>
<tr>
<td>License date</td>
<td>Feb 17, 2010</td>
</tr>
<tr>
<td>Licensed content publisher</td>
<td>John Wiley and Sons</td>
</tr>
<tr>
<td>Licensed content publication</td>
<td>Microwave and Optical Technology Letters</td>
</tr>
<tr>
<td>Licensed content title</td>
<td>A tunable bandpass filter based on varactor loaded split-ring resonators</td>
</tr>
<tr>
<td>Licensed content author</td>
<td>Genc Alper, Baktur Reyhan</td>
</tr>
<tr>
<td>Licensed content date</td>
<td>Jul 23, 2009</td>
</tr>
<tr>
<td>Start page</td>
<td>2394</td>
</tr>
<tr>
<td>End page</td>
<td>2396</td>
</tr>
<tr>
<td>Type of use</td>
<td>Dissertation/Thesis</td>
</tr>
<tr>
<td>Requestor type</td>
<td>Author of this Wiley article</td>
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<tr>
<td>Format</td>
<td>Print and electronic</td>
</tr>
<tr>
<td>Portion</td>
<td>Full article</td>
</tr>
<tr>
<td>Will you be translating?</td>
<td>No</td>
</tr>
<tr>
<td>Order reference number</td>
<td></td>
</tr>
<tr>
<td>Total</td>
<td>0.00 USD</td>
</tr>
</tbody>
</table>

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