2018

Novel Topologies Based Rf Filtering Components And Methodologies For Wireless Communication System

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NOVEL TOPOLOGIES BASED RF FILTERING COMPONENTS AND METHODOLOGIES FOR WIRELESS COMMUNICATION SYSTEM

by

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Submitted in Partial Fulfillment of the Requirements
For the Degree of Doctor of Philosophy in
Electrical Engineering
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2018

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DEDICATION

To my beloved family
ACKNOWLEDGEMENTS

The four-year Ph.D. experience in the University of South Carolina is an unforgettable journey filled with ups and downs, through which a brand-new future opens its door to me. It helps the starting point of my long career reach a new height. During my Ph.D. study, I received all kinds of help from people around me, and I couldn't have finished this dissertation without the support from them. It is my great pleasure to acknowledge them for their guidance, assistance and company.

Foremost, I would like to sincerely express my gratitude to my adviser Dr. Guoan Wang. Without him offering the second opportunity to let me participate in his research group and study under his guidance, I would have already quitted my Ph.D. study at Oregon State University and gone back to China. It is him that changes my life trajectory. Until now, I'm still feeling grateful to Dr. Wang that I could become a member of his SMART Lab four years ago. His inspiring ideas, broad scientific knowledge, and deep technology insight impress me all the time, and lead me in all the aspects of my research work. Dr. Wang gives me sufficient freedom in my research to work with the topics I'm interested in. Whenever I have difficulties in research, he is always there to provide enthusiastic encouragement and patient guidance, and he is always able to come up with smart and efficient methods to tackle the issues. But what makes me feel touched is his generous support when I have difficulties in life. From my heart, he is not only my adviser, but also my lifelong friend. In addition to scientific knowledge, what I have learned most from Dr. Wang is his optimistic attitude towards life, which will influence
me and encourage me against frustrations and failures throughout my life. Honestly, I
could not have imagined having a better adviser than him.

I would like to thank Prof. Mohammod Ali, Prof. Grigory Simin and Prof. Chen Li for their valuable time and effort serving in my Ph.D. committee and for their brilliant, constructive and precious suggestions in the proposal and dissertation defense. I'm very grateful to the principle engineer Yanzhu (Eric) Zhao from Medtronic Inc. and the principle engineer Xiaoding (Kevin) Cai from Cisco Inc. for mentoring me during my internship.

I am indebted to my colleagues and friends in the SMART Lab: Yujia Peng, Tengxing Wang, and Yongmao Huang, for stimulating research discussions, for the great help in simulation, fabrication and measurement, and for all the progress we have made together under the guidance of Dr. Guoan Wang. I also want to thank my office mates: Wuzhao Yan, Zhichao Liu, Zheqing Zhou, Lixing Yang, Guangxing Niu and Shijie Tang, for their friendship and company. Because of them, my Ph.D. life is rich, colorful, and memorable. Many thanks to the friends who helped me integrate into the overseas school life and make it memorable.

Most importantly, I would like to express my eternal love and appreciation to my beloved parents, Shiwu Jiang and Yonghong Xuan. Thank you very much for your endless love, unlimited support and understanding. I'm proud of being your son, and hope I could be your pride.

I reserve this final paragraph for my beautiful and great wife Na Han. I know these four years of not being together are too much for you. I could not find any single
word to express my deepest gratitude to you for raising our lovely son Nan Jiang independently in China.
ABSTRACT

Driven by the rapid progress of wireless communication technology in the past several decades, multiple generations of cellular technologies have been developed, deployed, and adopted to provide more convenient communication services to users. Nowadays, the personal hand-held devices, supporting multiple wireless standards, have been a multimedia terminal encompassing elements and functions such as video callers, Internet connectivity, home appliances remote controller, GPS, TV reception, and beyond. In order to accommodate a variety of wireless standards in a single device without imposing a substantial increase in cost and size, current and future RF transceiver front-ends should be designed with more attention. The main objective of this dissertation is to study new design topologies and implement a series of high performance RF filtering components which play critical roles in miniaturized RF transceivers supporting multiple wireless standards. A compact dual-band filter with high selectivity and wide rejection band, a filtering Wilkinson power divider, and balanced filters with fixed/reconfigurable center frequencies are proposed and successfully developed. In addition, an equation-based methodology is also first proposed and fully investigated to realize high-level integration of multiple functional modules in a single wireless device.

In the first part of this dissertation, a fourth-order cross coupling topology is first studied and applied to the design of a microstrip single-band filter and a microstrip dual-band filter. Along with the 0° feed structure, a total number of four finite transmission
zeros are generated locating close to their passbands, thus greatly improving the frequency selectivity of the designed filters. The rejection band of the single-band filter is greatly extended with a lowpass unit added to the input/output feed structure. The strategy of staggering harmonic peaks of occupied resonators is also adopted in the dual-band filter design for the purpose of comparison, and considerable size miniaturization and insertion loss reduction have been achieved.

The second part of this dissertation is focused on the development of a filtering Wilkinson power divider. Two coupled T-stub loaded dual-mode resonators are used to replace the quarter-wavelength line segments in order to provide filtering response for the Wilkinson power divider. Due to the mixed electric and magnetic coupling between two coupled T-stub loaded dual-mode resonators, an inter-cross-coupled topology is created, thus generating a total number of four transmission zeros near the operation band, which provides good frequency selecting characteristic for the developed power dividing component.

The third part of this dissertation mainly discusses and solves crosstalk issues brought by the high density integration. Solutions are provided in the design of both component level and system level. On the component level, RF dual-band filters with fixed/reconfigurable center frequencies in the balanced architecture are proposed and designed with higher immunity to electromagnetic interference and harmful environmental noise. On the system level, an equation based methodology is developed for the layout optimization of closely spaced microstrip traces with tab routing, which mitigates the crosstalk issue raised by integrating multiple functional modules with parallel routing traces in a highly compact manner. With this useful methodology
available, the overall performance of highly integrated wireless communication system will be improved.
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CHAPTER 1
INTRODUCTION

1.1 Motivation and Background

The past several decades have witnessed the development, deployment, and adoption of multiple generations of cellular technologies, from a single-mode, triple-band 2G system to a triple-mode, 9-band (4×GSM, 5×UMTS with HSPA+) high-speed data-capable system in year 2010. Driven by the rapid progress of wireless communication technology, the cellular technologies have been improved by several orders of magnitude in terms of data rates and general capabilities, as clearly shown in Fig. 1.1. In the upcoming 5G communication technology, the trend of supporting multi-mode and multi-standard operation will continue to ensure higher data transmission rates and increased functionality for various applications. Fig. 1.2 illustrates such a scenario that several wireless standards are compatibly supported by a modern cellphone, which has evolved into a complex product that goes beyond being a mere communication device. It has become a multimedia terminal encompassing elements and functions such as Internet connectivity, home appliances controller, TV reception, video callers, and so on.

With no doubt, the high-level integration of different functional modules supporting each individual communication standard will have improvements in reducing the size and component count of today’s complex wireless communication systems. However, it will also raise significant challenges to the RF designers, since the electromagnetic interference from nearby functional modules could easily distort the...
performance of basic RF components, thus degrading the whole system performance. In order to facilitate the demand of implementing multiple standards in a single wireless device without sacrificing the performance of whole system and imposing a substantial cost and size, RF components constituting transceiver front-ends should be designed with more attention. So far, a lot of research efforts have been endeavored on this topic and
various approaches developed can be classified into the following categories, as detailed below:

The first category is to design individual RF passive components capable of manipulating different types of RF signals at various frequencies. Fig. 1.3 depicts the system architecture of a highly integrated RF transceiver front-end system that supports all 2G/3G/4G modes. Separate RF front-ends are parallel integrated together to support different frequency bands of multiple wireless standards. According to [1], off-chip passives actually occupy 80% of the total transceiver board area and account for 70% of

Figure 1.3 System architecture of a RF front-end system supporting all 2G/3G/4G Modes.
the cost. As a key building block in RF front-ends, bandpass filter exists within each separate RF front-end, as demonstrated in Fig. 1.3. If one bandpass filter could support multi-band operation and meet the divergent specifications at different standards such as operation frequency, modulation bandwidth, and power level, the total amount of required RF filters could be reduced.

The second category is to integrate two or more separate circuit topologies into a single RF component. It is known to all that the conventional RF system usually employs discrete devices to implement individual function, which leads to bulky size and large

![Figure 1.4 Cascaded filter and power divider and filtering power divider [2].](image)

![Figure 1.5 Illustration of the concept of tunable RF front-end.](image)
insertion loss. As a promising solution, integrating multiple functions into one component is an effective method for size and cost reduction. For instance, if the power divider and filter coexist in the same front-end system, they can be merged together as shown in Fig. 1.4 [2], resulting in compact size and low loss.

The third approach is to look for technological innovations from the system level. In recent years, the concepts such as software defined radio (SDR) and cognitive radio (CR) have been introduced and widely employed [3], [4]. Fig. 1.5 shows the concept of a tunable RF system. Controlled by a variety of outer tuning mechanisms, the hardware parameters could be adapted to the requests of different wireless standards. The advantages of such reconfigurable RF systems include efficient usage of limited electromagnetic spectrum and complexity reduction of modern wireless communication system. To accomplish the design of miniaturized, frequency-agile, and multi-functional systems, tunable RF components are apparently indispensable. For example, a large amount of research outcomes have been published regarding the replacement of switched banks of static filters in Fig. 1.3 by tunable filters [5]-[8].

Last but not the least, balanced components known to have higher immunity to electromagnetic interference and environmental noises as well as good dynamic range are increasingly adopted to develop fully differential RF front-end architectures. The interference signal easily distorting the operation of single-ended RF components will be added to each of the two balanced signals equally, as shown in Fig. 1.6. Thus, the error will be canceled out, making differential signal chains less susceptible to interference. Furthermore, under the trend of system-in-package and system-on-chip, high density signal routings should also be considered with caution because of the far-end crosstalk
due to electromagnetic coupling. It is believed that the performance degradation of whole wireless communication system caused by high level integration could be effectively reduced by the mitigation of far-end crosstalk coming from routing wires.

All in all, the approaches mentioned above basically conclude the endeavors to reduce the cost and size of complex RF front-end systems supporting multi-mode/multi-standard operation. With those in mind, this dissertation is focused on the development of a series of high performance RF filtering components based on novel topologies including a compact dual-band filter with high selectivity and wide rejection band, a filtering Wilkinson power divider, and balanced dual-band filters with fixed/reconfigurable center frequencies. In addition, an equation-based methodology is also investigated for the purpose of realizing high-level integration of multiple functional modules in a single wireless device.

Figure 1.6 Comparisons between (a) unbalanced wiring and (b) balanced wiring under the influence of interference.
1.2 Overview of Past Relevant Research

The rapidly developing wireless communication technology and the increasingly emerging applications are always challenging filter designers with more stringent performance requirements, such as multi-band operation, multi-functionality, and flexible reconfigurability. It is relatively straightforward to accomplish the design of a simple bandpass filter with certain specifications. However, how to implement multi-band operation, multi-functionality and flexible reconfigurability still remains a topic worth of research attention and efforts. In the following section, past relevant research in those areas will be reviewed and briefly discussed.

1.2.1 Multi-Band Bandpass Filters

The development of a number of wireless communication standards imposes a requirement for multi-band RF filters that work at two or more frequencies corresponding to different standards like IEEE 806.16, WIFI, Bluetooth, GSM and CDMA. Till now, a lot of dual-band bandpass filters have been published [9]-[16] and the design methods could be summarized as follows,

(1) The simplest method for multi-band filter design is to cascade a wideband bandpass filter and a bandstop filter, as presented in [9]. Fig. 1.7 shows the fabricated circuits. Despite that this method is intuitive and straightforward for implementing dual-band filter, the cascade connection inevitably leads to greater circuit size and large insertion loss. In [10] and [11], some improved designs are proposed by comprising the individual bandpass filter and bandstop filter together. For example, the schematic of the dual-band filter published in [11] is shown in Fig. 1.8. Short-circuited quarter-wavelength
shunt stubs separated by quarter-wavelength transmission lines will generate the bandpass response while serial LC circuits between adjacent quarter-wavelength stubs produce a bandstop response whose bandwidth is made narrower than passband bandwidth. As a matter of fact, these dual-band filters [9]-[11] are all based on the same idea although implemented in different forms. The characteristics of each passband could be managed by adjusting the characteristics of both the bandpass and bandstop filter.

Figure 1.7 Fabricated (a) Chebyshev type-II bandpass filter, (b) Butterworth bandstop filter as well as (c) dual-band filter consisting of a bandpass filter and a bandstop filter in cascade connection, presented in [9].
(2) Another popular way of achieving dual-band response is to combine two sets of single-band filters [12], [13]. The design procedures of such kind of dual-band filter include individual design of two single-band filters and subsequent combination of them with common input/output. Once combined together, one passband performance may be influenced by the other. As a result, a final optimization is needed to obtain the desired dual-band performance. It is worth mentioning that the building blocks of each single-band filter could be of the same type, as shown in Fig. 1.9, or of different types, as shown in Fig.

![Figure 1.8 Schematic of the dual-band filter in [11].](image)

![Figure 1.9 Schematic of the dual-band filter in [12].](image)
By this approach, the specifications of each passband such as working frequency and bandwidth can be individually manipulated. This method provides great design freedom in specifying each passband performance of dual-band filters. The disadvantage is still relatively large circuit size occupied by multiple resonators.

(3) Last, the dual-band behavior of RF filters can be constructed from the fundamental frequency and the controllable first harmonic of stepped impedance resonators [14], [15] or stub-loaded multi-mode resonators [16]. The detailed configuration of stepped impedance resonator (SIR) enabled dual-band filter is exhibited in Fig. 1.11 [14]. The ratio between the first harmonic and the fundamental mode is manipulated by the impedance ratio and length ratio, as detailed in [15]. Apart from SIR, stub-loaded resonator

![Figure 1.10 Layout of the dual-band filter in [13].](image-url)
also provide dual-mode behavior and the first dual-band filter employing SLR is given in Fig. 1.12 [16]. Due to its symmetry, odd- and even-mode is applied for analysis, as described in [16]. It is widely acknowledged that adopting dual-mode resonators for dual-band filter design is beneficial for miniaturization but it lacks the flexibility to meet the specifications at both passbands simultaneously.

Figure 1.11 Illustration of SIR enabled dual-band filter in [14].

Figure 1.12 A Dual-band filter employing SLRs in [16].

1.2.2 Filtering Function Integrated RF Components

RF Passive components like filter, antenna, coupler, and power divider are usually larger components compared to other components in the RF front-ends of modern wireless communication systems. They occupy a considerably large circuit size on the
transceiver front-ends board. Thus, it will be of great interest if a single compact module can fuse multiple functions for size reduction. In the past several years, some integrated designs are proposed, for instance, filter-antenna [17]-[19], filter-coupler [20], [21], and filter-power divider [22]-[24]. A typical example from each group will be briefly introduced in the following section.

![Figure 1.13 Integrated three-pole filter-antenna in [17].](image)

In [17], a new compact filter-antenna for modern wireless communication systems is proposed, as demonstrated in Fig. 1.13. As observed, two microstrip square open-loop resonators, a coupled line, an a $\Gamma$-shaped antenna are used and integrated to constitute a filter-antenna. The transmission loss between the filter and antenna is reduced to zero, making the total loss of filter-antenna almost identical to the filter insertion loss alone. The electrical response of the filter-antenna presents integrated filtering and radiating functions simultaneously.
Fig. 1.14 illustrates the microstrip layout of a filtering 180° hybrid coupler, which is developed in [20]. Based on four coupled half-wavelength resonators, the proposed filtering 180° hybrid coupler can provide power division and phase shift together with a second-order bandpass response to suppress the spurious frequencies or the intermodulation products generated by amplifiers.

As an example of filter-power, Fig. 1.15 exhibits a bandpass power divider with improved skirt selectivity [22]. The conventional Wilkinson power divider is composed of two quarter-wavelength sections at designed frequency, leading to a large circuit size. In this design, the quarter-wavelength sections are replaced by four compact cross-coupled net-type resonators. This kind of resonator-based filtering power divider is able to realize in-band equal power splitting as well as quasi-elliptic bandpass response.

In summary, integrating multiple functionalities into a single component is beneficial for reducing circuit number and/or space of RF/microwave system as well as manufacturing cost.
1.2.3 Reconfigurable Bandpass Filters

Utilizing reconfigurable RF front-ends is an interesting way to provide diverse services to cater for the requirements of multi-mode, multi-standard operation. As one of

Figure 1.15 Schematic layout of a quasi-elliptic bandpass power divider in [22].

Figure 1.16 Tuning screws enabled tunable filter in [26]: (a) schematic and (b) photograph of fabricated real sample.
the key passive RF/wireless components, high performance tunable filters can result in better channel selectivity, decreased complexity, and reduced circuit dimension. Up to now, a variety of tuning mechanisms have been applied to design tunable RF filters [25]-[37], as detailed below:

1. Mechanically tunable technique enabled by shifting a material or tuning screws is relatively early technology to implement tunability [25]-[27]. In [25], the dielectric slab can be moved vertically above a RF filter, or used to

Figure 1.17 (a) Scanning electron microscopy (SEM) view. (b) Photograph view of a fabricated MEMS tunable filter in [28].
generate deformation on a conductive film to tune working frequency of
dielectric resonator filters or evanescent-mode cavity filters. Fig. 1.16
demonstrates an example of adopting tuning screws to enable tuning
capability of filter center frequency, as introduced in [26]. The mechanical
tuning technique can provide high-Q and high power-handling capabilities,
but suffer from low tuning speed, bulky size, heavy weight, and integration
issue with other printed circuits.

(2) To overcome the size and integration issue of conventional mechanical tuning
mechanism, Microelectromechanical Systems (MEMS) has been successfully
applied in tunable RF device topologies [28]-[30]. Fig. 1.17 shows the SEM
picture and photograph of a fabricated continuously tunable MEMS bandpass
filter implemented in a third-order coupled resonator configuration [28].
Continuous electrostatic tunability is achieved by three tunable capacitor
banks. The good merits of tunable filters applying RF MEMS technology

![Figure 1.18](image)

**Figure 1.18** (a) The topology and (b) Fabricated tunable BPF with BST varactor chips in [31].
include low loss, high linearity, and low power consumption. The limitations of MEMS technology are high response time, complicated fabrication and low reliability.

(3) In addition to aforementioned tuning techniques, functional materials provide another tuning strategy for tunable filter applications [31], [32]. Barium Strontium Titanate (BST) and Lead Zirconate Titanate (PZT) belong to the category of ferroelectric material, whose materials’ permittivity is subject to change with the applied outer electric filed. As an example, Fig. 18 presents a tunable dual-mode filter, in which the BST interdigital varactors are deployed as tuning elements [31]. Apart from ferroelectric materials with tunable permittivity, ferromagnetic materials are known to have adjustable permeability with external biasing magnetic field [33], [34]. Yttrium-iron-garnet (YIG) is a kind of ferromagnetic material and could be used as the substrate for tunable bandpass filter, when partially magnetized [33], as given in Fig. 1.19. As external biasing magnetic field is provided, the permeability

![Figure 1.19 Geometry of the tunable BPF fabricated on partially magnetized YIG [33].](image-url)
of YIG substrate can be changed. Overall, these functional materials suffer from considerable material loss, non-linearity and temperature sensitivity and most importantly, they rely on outer bias, which introduces the integration issue.

(4) Semiconductor varactor diodes, such as PIN diodes and GaAs Schottky diodes are believed as the most popular tuning elements for designing tunable RF filters. They are of particular interest thanks to their availability, low cost, high capacitance tuning range, and low response time. As a result, a great deal of work has been done with respect to tunable filters based on semiconductor

![Figure 1.20 (a) Schematic and (b) stripline-realization of a 4-pole tunable bandpass filter in [35].](image)
varactor diodes [35]-[37]. In [35], a 4-pole stripline bandpass filter with tunable center frequency and bandwidth is demonstrated, as shown in Fig. 1.20. By flexibly controlling varactor diodes, the intrinsic transmission zeros can be moved from the upper stopband to the lower stopband according to specific application demands. Furthermore, it could also result in a zero-coupling filter with high isolation and near absorptive-filter performance. However, the major drawbacks of semiconductor varactors are low quality factor, low power handling capability, low linearity, and auxiliary biasing networks are required, thus inevitably increasing the complexity of systems.

1.3 Chapter Outline

Motivated by incredibly high demands for high performance RF components in current wireless communication technology and inspired by the literature review demonstrated in previous sections, the primary effort of this research is focused on developing multi-band bandpass filters, filtering function integrated passive components, balanced RF filters as well as reconfigurable RF filters, which are of great importance in practical applications since multiple standards are simultaneously supported by single mobile device. In addition, an equation-based methodology is proposed to optimize the closely spaced microstrip traces with tab routing for crosstalk mitigation, under the trend of high-density integration. Accordingly, this dissertation is organized as follows:

Chapter 2 quickly overviews two traditional filter design theories, namely, image-parameter theory and insertion-loss theory, respectively. These two design theories developed in early twentieth century are adequate for the technology and demands at that time. Since late twentieth century, elliptic or quasi elliptic RF filters with finite transmission zeros in the transmission response are becoming popular and they calls for novel design technique,
coupling matrix theory, which will be subsequently reviewed in detail. After that, the generation mechanisms of transmission zeros are illustrated with coupling matrix theory.

Chapter 3 will introduce the design of fourth-order cross coupled microstrip single-band filter and dual-band filter with high frequency selectivity and wide rejection band. The fourth-order cross coupled topology for filter design is investigated by coupling matrix theory. Along with 0° feed structure, up to two pairs of finite transmission zeros are locating close to the passband and will contribute to improved frequency selectivity. For extending the rejection band of the single-band filter, a lowpass unit is added to the input/output feed structure. However, the method of inserting extra structure to suppress the harmonic peaks leads to larger circuit size and also increased insertion loss from input to output. In the dual-band filter design, stub-loaded uniform-impedance resonators are replaced by stub-loaded stepped-impedance counterparts. By staggering harmonic peaks of occupied resonators, rejection band of the dual-band filter is extended intrinsically, without introducing any extra structures. Two filter samples are fabricated with standard PCB technology for verification purpose. The agreements between theoretical and experimental results are obtained.

Chapter 4 will talk about the multi-functional RF component that fuses the functions of multiple single RF components. The functions of frequency selection and power division are integrated into the proposed single device. Two coupled dual-mode resonators incorporating mixed electric and magnetic coupling are used to replace the quarter-wavelength line segments of conventional Wilkinson power divider. The proposed filtering power divider can perform not only the function of power splitting but also the function of frequency selecting. The multi-functional component introduced in
this chapter is compact in size and the measured results of fabricated sample show consistency with the simulated ones.

Chapter 5 will discuss two RF dual-band filters in balanced architecture. Among them, one is a dual-band balanced filter with fixed center frequency, and the other is a dual-band balanced filter with reconfigurable center frequencies. As known, the balanced filter design has one more requirement than single-ended filter, which is common-mode suppression within the differential-mode passbands. With that requirement taken into consideration, relatively simple second-order coupling topology is adopted. Novel dual-stub loaded resonators are presented in this chapter to provide dual-mode resonances at two diverse frequencies under differential-mode excitation. To enable frequency tuning capabilities, commercial varactor diodes are embedded into the dual-stub loaded resonators. The experimental results all match the theoretical expectations very well.

Chapter 6 will focus on developing an equation-based methodology for optimizing the dimension of closely spaced microstrip traces with tab routing. The introduction of tab routing into microstrip traces is reported by Intel Corporation to mitigate the far-end crosstalk. Except for time-consuming numerical simulation, no analytical solution with acceptable accuracy could be found to predict the S-Parameters of microstrip tabbed routing. In this chapter, an accurate equation-based methodology capable of predicting S-Parameters is proposed. Analytical results are compared with numerical counterparts for confirming its accuracy. With this useful methodology, the overall performance of highly integrated wireless communication system will be improved by optimizing the high density routing traces.
Chapters 7 firstly gives a summary of contributions to the dissertation and then presents future work to be performed.
CHAPTER 2
FILTER DESIGN THEORIES AND RF RESONATORS

2.1 Origin of Filters and Traditional Filter Design Theory

Nowadays, with the advancement of more and more electric systems like GPS, satellite communication, radar and mobile communication, finite electromagnetic spectrum is being heavily occupied to accommodate as many systems as possible. Inevitably, the assigned adjacent channels are quite close to each other. As indispensable components in these systems, filters are playing a much more crucial role than ever. As a matter of fact, filter in circuit theory is a kind of electrical network that is capable of manipulating the amplitude and/or phase characteristics of a signal with respect to its various frequency components. Described in frequency domain, a filter would allow the signals of desired frequencies pass through and block the signals of unwanted frequencies. The design theory of filter should date back to early twentieth century, when in the United States and Germany, Campbell and Wagner independently concluded the basic concepts of filter based on the knowledge of loaded transmission line and the classical theory of vibration systems. Stemming from Campbell and Wagner’s pioneering work, the filter design theory has gradually developed towards two directions, thus leading to the formation of image-parameter theory and insertion-loss theory for filter design [38].
The image-parameter theory was proposed in the late 1930s and was useful for low-frequency filters in radio and telephony. Since their working frequencies are typically below 1 GHz, lumped elements such as resistors, inductors and capacitors are employed to design filters with the image-parameter method, which divides a filter into a cascade of simpler two-port sub-networks, and then attempt to come up with the schematic of each two-port sub-network. Once combined, they will provide the desired cutoff frequencies and attenuation characteristics. However, the drawback of image-parameter approach is that it does not allow the specification of a particular frequency response over the complete operating range. In other words, the frequency response of filter designed by image-parameter method cannot be arbitrarily shaped. Therefore, it could only provide a loose control over the characteristics of passband and stopband.

In contrast to the image-parameter theory, the insertion-loss theory is more powerful in the sense that it could provide the specified response of the filter. It starts with a complete specification of a physically realizable frequency characteristic by transfer function, from which a suitable filter schematic will be synthesized using network synthesis techniques to fulfill the transfer function. The design procedure is simplified by beginning with lowpass filter prototypes normalized in terms of impedance and frequency. Subsequently, frequency scaling and impedance transformation are applied to denormalize the lowpass prototype and synthesize different type of filters with different cutoff frequencies.

It is worth mentioning that both the image-parameter theory and the insertion-loss theory of filter design lead to circuits using lumped elements. To cater for high frequency applications, distributed elements consisting of transmission line sections are employed.
instead of lumped elements with aid of Richard's transformation and Kuroda's identities [39]. All in all, the above-mentioned filter design methods involving extraction of the electrical elements (for instance: inductors, capacitors as well as transmission line sections) from polynomials mathematically describing filter response, are adequate for the technology and demands belong to the early twentieth century. Currently, filters exhibiting the Chebyshev and Butterworth response are already not sufficient to meet the stringent demands of high quality factor, constant group delay, low in-band insertion loss, and high out-of-band suppression. The filter design has consequently evolved from the utilization of lumped LC resonator to microstrip resonator [40], [41], waveguide resonator [42], [43], dielectric resonator [44], substrate integrated waveguide resonator [45], [46] and surface/bulk acoustic resonator [47], [48]. To prevent signal interference from the adjacent channels, elliptic or quasi-elliptic filters known to possess finite transmission zeros at desired frequencies are becoming increasingly suitable. Therefore, novel filter design techniques are developed as well to assist designers in achieving high performance filters with finite transmission zeros close to the passband.

2.2 Coupling Matrix for Determining S-Parameters

After a thorough literature search [49]-[54], the most widely adopted transmission zero generation is introducing extra cross coupling mechanism in the design of coupled-resonator filters that can enable multi-path signal transmission between input and output ports. At certain frequencies, signals along different paths are of same amplitude but opposite phase and they can cancel each other, thus generating transmission zeros. The general design method of coupled-resonator filters is based on the coupling coefficients
between intercoupled resonators and the external quality factors of input/output resonators, by which the positions of transmission zeros are uniquely determined as well.

The generalized equivalent circuit model of a bandpass filter consisting of $N$-coupled resonators with cross-coupling incorporated is shown in Fig. 2.1. The frequency band of interest is assumed to be narrow. Each resonator under analysis is represented by a lossless $LC$-resonator and $R_i/R_N$ is the resistance termination at input/output, respectively. In addition, $i_i$ denotes the loop current in the $i$th closed network and $e_s$ represents for the voltage source. According to Kirchhoff's voltage law, the loop equations for $N$ closed networks are concluded as follows [55]:

\[
\begin{align*}
\left( j\omega L_1 + \frac{1}{j\omega C_1} + R_1 \right) i_1 - j\omega L_{12} i_2 \cdots - j\omega L_{1N} i_N &= e_s \\
-j\omega L_{21} i_1 + \left( j\omega L_2 + \frac{1}{j\omega C_2} \right) i_2 \cdots - j\omega L_{2N} i_N &= 0 \\
\vdots \\
-j\omega L_{N1} i_1 - j\omega L_{N2} i_2 \cdots + \left( j\omega L_N + \frac{1}{j\omega C_N} + R_N \right) i_N &= 0
\end{align*}
\] (2.1)

This set of equations is equivalent to the following expression in a matrix form:
\[
[Z]_{N \times N} \begin{bmatrix} i_1 \\ i_2 \\ \vdots \\ i_N \end{bmatrix} = \begin{bmatrix} e_s \\ 0 \\ \vdots \\ 0 \end{bmatrix}
\] (2.2)

where \( N \times N \) impedance matrix \( [Z] \) is expressed by

\[
[Z] = \begin{bmatrix}
  j\omega L_1 + \frac{1}{j\omega C_1} + R_1 & -j\omega L_{12} & \cdots & -j\omega L_{1N} \\
  -j\omega L_{21} & j\omega L_2 + \frac{1}{j\omega C_2} & \cdots & j\omega L_{2N} \\
  \vdots & \vdots & \ddots & \vdots \\
  -j\omega L_{N1} & -j\omega L_{N2} & \cdots & j\omega L_N + \frac{1}{j\omega C_N} + R_N 
\end{bmatrix}
\] (2.3)

Accordingly, a new normalized impedance matrix \( \tilde{Z} \) is defined as

\[
\tilde{Z} = [Z]_{N \times N} / (\omega_0 L \cdot FBW)
\] (2.4)

\[
\tilde{Z} = \begin{bmatrix}
  p + \frac{R_1}{\omega_0 L \cdot FBW} & -j \frac{\omega}{\omega_0} M_{12} & \cdots & -j \frac{\omega}{\omega_0} M_{1N} \\
  -j \frac{\omega}{\omega_0} M_{21} & p & \cdots & -j \frac{\omega}{\omega_0} M_{2N} \\
  \vdots & \vdots & \ddots & \vdots \\
  -j \frac{\omega}{\omega_0} M_{N1} & -j \frac{\omega}{\omega_0} M_{N2} & \cdots & p + \frac{R_N}{\omega_0 L \cdot FBW} 
\end{bmatrix}
\] (2.5)

where \( p = j \frac{\omega_0}{\Delta \omega} \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) \), and \( M_{ij} = \frac{L_{ij}}{L} \). It is worth mentioning that in a synchronously tuned filter, \( L = L_1 = L_2 = \cdots = L_N \). The above equation is simplified by defining scaled external quality factor and normalized coupling coefficient as,

\[
\frac{1}{q_{e1}} = \frac{R_1}{\omega_0 L \cdot FBW}, \quad \frac{1}{q_{e1N}} = \frac{R_N}{\omega_0 L \cdot FBW}
\] (2.6)

\[
m_{ij} = \frac{M_{ij}}{FBW}
\] (2.7)

Subsequently, the matrix \( \tilde{Z} \) is re-written by

\[
\tilde{Z} = \begin{bmatrix}
  p + \frac{1}{q_{e1}} & -jm_{12} & \cdots & -jm_{1N} \\
  -jm_{21} & p & \cdots & -jm_{2N} \\
  \vdots & \vdots & \ddots & \vdots \\
  -jm_{N1} & -jm_{N2} & \cdots & p + \frac{1}{q_{eN}} 
\end{bmatrix}
\] (2.8)

As a consequence, the loop current \( i_1 \) and \( i_N \) are solved by
\[
\begin{align*}
i_1 &= \frac{1}{\omega_0 L \cdot FBW} \begin{bmatrix}
p + \frac{1}{q_{e1}} & -jm_{12} & \cdots & -jm_{1N} \\
-jm_{21} & p & \cdots & -jm_{2N} \\
\vdots & \vdots & \ddots & \vdots \\
-jm_{N1} & -jm_{N2} & \cdots & p + \frac{1}{q_{eN}}
\end{bmatrix}^{-1} \cdot e_s 
\end{align*}
\]

\[
\begin{align*}
i_N &= \frac{1}{\omega_0 L \cdot FBW} \begin{bmatrix}
p + \frac{1}{q_{e1}} & -jm_{12} & \cdots & -jm_{1N} \\
-jm_{21} & p & \cdots & -jm_{2N} \\
\vdots & \vdots & \ddots & \vdots \\
-jm_{N1} & -jm_{N2} & \cdots & p + \frac{1}{q_{eN}}
\end{bmatrix}^{-1} \cdot e_s 
\end{align*}
\]

If the generalized model shown in Fig. 2.1 is considered as a two-port network, the incident and reflection variables at two ports are described by \(a_1, a_2\) and \(b_1, b_2\), respectively. They are related with \(V_1, I_1\) and \(V_2, I_2\) by the following equations [55],

\[
\begin{align*}
a_n &= \frac{1}{2} \left( \frac{v_n}{\sqrt{R_n}} + \sqrt{R_n} i_n \right) \\
b_n &= \frac{1}{2} \left( \frac{v_n}{\sqrt{R_n}} - \sqrt{R_n} i_n \right)
\end{align*}
\]

for \(n = 1\) and 2

Besides that, it is inspected and identified that \(I_1 = i_1, V_1 = e_s - i_1 \cdot R_1\), and \(I_2 = -i_N, V_2 = i_N R_N\). Therefore,

\[
a_1 = \frac{1}{2} \frac{e_s}{\sqrt{R_1}}, a_2 = 0 \quad (2.12)
\]

\[
b_1 = \frac{e_s - 2i_1 R_1}{2\sqrt{R_1}}, b_2 = i_N \sqrt{R_N} \quad (2.13)
\]

By the definition of S-parameters [39], we have

\[
S_{11} = \frac{b_1}{a_1} \bigg|_{a_2=0} = 1 - \frac{2R_1 i_1}{e_s} \quad (2.14)
\]

\[
S_{21} = \frac{b_2}{a_1} \bigg|_{a_2=0} = \frac{2\sqrt{R_1 R_N i_N}}{e_s} \quad (2.15)
\]

Plugging Eq. 2.9 and Eq. 2.10 into Eq. 2.14 and Eq. 2.15 leads to
\[ S_{11} = 1 - \frac{2R_1}{\omega_0 L \cdot FBW} \begin{bmatrix} p + \frac{1}{q_{e1}} & -jm_{12} & \cdots & -jm_{1N} \\ -jm_{21} & p & \cdots & -jm_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ -jm_{N1} & -jm_{N2} & \cdots & p + \frac{1}{q_{eN}} \end{bmatrix}^{-1} \]  

(2.16)

\[ S_{21} = \frac{2 \sqrt{R_1 R_N}}{\omega_0 L \cdot FBW} \begin{bmatrix} p + \frac{1}{q_{e1}} & -jm_{12} & \cdots & -jm_{1N} \\ -jm_{21} & p & \cdots & -jm_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ -jm_{N1} & -jm_{N2} & \cdots & p + \frac{1}{q_{eN}} \end{bmatrix}^{-1} \]  

(2.17)

To sum up, the S-parameters of a given filter can be determined by its coupling coefficients through above derived formulae. However, the coupling between source/load and resonators are not taken into consideration at all. To overcome that shortcoming, a more general form of coupling matrix is proposed by S. Amari and R. J. Cameron [56]-[58] for the development of a more complete filter design theory based on coupling matrix. Fig. 2.2 illustrates the schematic layout of \( N \)-coupled resonators with multiple source/load coupling. Its corresponding \( (N+2) \times (N+2) \) coupling matrix \([m]\) includes not only the main coupling information between resonators, but also extra input/output
couplings from source/load terminations to resonators as well as direct coupling between source and load. The filter response in terms of scattering parameters are computed by,

\[ S_{11} = 1 + 2j[A]^{-1}_{11} \]  
\[ S_{21} = -2j[A]^{-1}_{N+2,1} \]  
\[ [A] = [m] + \Omega[U] - j[R] \]

\[ [m] = \begin{bmatrix} m_{SS} & m_{S1} & m_{S2} & \cdots & m_{SN} & m_{SL} \\ m_{S1} & m_{11} & m_{12} & \cdots & m_{1N} & m_{1L} \\ m_{S2} & m_{12} & m_{22} & \cdots & m_{2N} & m_{2L} \\ \vdots & \vdots & \vdots & \ddots & \vdots & \vdots \\ m_{SN} & m_{1N} & m_{2N} & \cdots & m_{NN} & m_{NL} \\ m_{SL} & m_{1L} & m_{2L} & \cdots & m_{NL} & m_{LL} \end{bmatrix} \]

where \( \Omega = \frac{\omega_0}{\Delta \omega} \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) \) is the frequency variable of lowpass prototype, \( U = \text{diag}[0, 1, 1, \cdots, 1, 0] \), and \( R = \text{diag} \left[ 1, \frac{1}{Q_1FBW}, \frac{1}{Q_2FBW}, \cdots, \frac{1}{Q_NFBW}, 1 \right] \).

2.3 RF Resonators

As mentioned at the beginning of Section 2.2, coupled-resonator filters are promising solutions for implementing elliptic or quasi-elliptic response with sharp roll-off skirt, and superior frequency selectivity. With no doubt, high performance RF resonators are indispensable and critical building blocks to construct these RF filters. Next, the resonance conditions of some RF resonators frequently used for coupled-resonator filter design are reviewed in the following section.

2.3.1 Half-Wavelength Resonator and Quarter-Wavelength Resonator

Fig. 2.3 depicts the schematic of an open-ended half-wavelength resonator and a quarter-wavelength resonator with one end shorted to ground. For simplicity, the losses are ignored in both cases. First, consider the half-wavelength resonator shown in Fig. 2.3
(a), the reflection coefficient $\Gamma = 1$ at position $x = 0$, and the voltage/current along the line are thus expressed by

$$V(z) = V_0^+ (e^{-j\beta z} + e^{j\beta z}) = 2V_0^+ \cos(\beta z)$$  \hspace{1cm} (2.22)$$

$$I(z) = \frac{V_0^+}{Z_0} (e^{-j\beta z} - e^{j\beta z}) = -\frac{2jV_0^+}{Z_0} \sin(\beta z)$$  \hspace{1cm} (2.23)$$

$$Z_{in} = -jZ_0 \cot(\beta z)$$  \hspace{1cm} (2.24)$$

Next, consider the quarter-wavelength resonator given in Fig. 2.3 (b), the reflection coefficient $\Gamma = -1$ at position $x = 0$, and the voltage/current along the line are similarly described by

$$V(z) = V_0^+ (e^{-j\beta z} - e^{j\beta z}) = -2jV_0^+ \sin(\beta z)$$  \hspace{1cm} (2.25)$$

$$I(z) = \frac{V_0^+}{Z_0} (e^{-j\beta z} + e^{j\beta z}) = \frac{2V_0^+}{Z_0} \cos(\beta z)$$  \hspace{1cm} (2.26)$$

$$Z_{in} = jZ_0 \tan(\beta z)$$  \hspace{1cm} (2.27)$$

Figure 2.3 (a) Half-wavelength resonator with open end and (b) quarter-wavelength resonator with shorted end.
According to the resonance condition $\text{Im}(Y_{in}) = 0$ [39], the resonance occurs when

\[ f_1 = \frac{c_0}{2L_1\sqrt{\varepsilon_{\text{eff}}}} \quad (2.28) \]

\[ f_2 = \frac{c_0}{4L_2\sqrt{\varepsilon_{\text{eff}}}} \quad (2.29) \]

where $f_1$ and $f_2$ represent for the resonant frequency of half-wavelength resonator and quarter-wavelength resonator, respectively. In addition, $c_0$ denotes the light velocity in free space and $\varepsilon_{\text{eff}}$ is the effective dielectric constant of surrounding media.

### 2.3.2 Metallic Waveguide Resonator

![Illustration of electric fields for TE101 and TE102 modes in a rectangular cavity resonator.](image)

Figure 2.4 Illustration of electric fields for TE$_{101}$ and TE$_{102}$ modes in a rectangular cavity resonator.

The schematic of a rectangular metallic waveguide resonator is shown in Fig. 2.4. It can be treated as a rectangular waveguide along $z$-direction and then short-circuited at position $z = 0$ and $d$. In addition to the boundary conditions required by a rectangular waveguide at the side walls ($x = 0, a$ and $y = 0, b$), we have to ensure that $E_x = E_y = 0$ at $z = 0, d$. Referring to [39], the wave number of TE$_{mnl}$ or TM$_{mnl}$ resonant mode of cavity resonator is defined by

\[ k_{mnl} = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2 + \left(\frac{l\pi}{d}\right)^2} \quad (2.30) \]

And the resonant frequency corresponding to TE$_{mnl}$ or TM$_{mnl}$ mode of cavity resonator is determined as
\[ f_{mn} = \frac{c_0 k_{mn}}{2 \pi \sqrt{\mu_r \varepsilon_r}} = \frac{c_0}{2 \pi \sqrt{\mu_r \varepsilon_r}} \sqrt{\left(\frac{m \pi}{a}\right)^2 + \left(\frac{n \pi}{b}\right)^2 + \left(\frac{l \pi}{d}\right)^2} \tag{2.31} \]

where \( m, n, l \) respectively stands for the number of variations in the standing wave pattern along \( x, y, z \) direction.

### 2.3.3 Substrate Integrated Waveguide Resonator

As plotted in Fig. 2.5, substrate integrated waveguide (SIW) is synthesized on planar substrates by periodically placing metallic via holes to connect the top and bottom metal plane [59]. With the existence of top and bottom metal plane as well as two pairs of via arrays, the EM waves are well confined. Due to the fact that longitudinal current flow does not exist on the side walls of SIW, TM modes are not supported by SIW. The height of SIW is determined by substrate thickness, which is usually much smaller than the SIW width. As a result, only TE\(_{m0}\) modes are of interest while utilizing SIW to design RF components.

Similar with the formation of rectangular waveguide resonator, a substrate integrated waveguide cavity resonator can be obtained by adding two more metallic via holes, as depicted in Fig. 2.6. The simple static equivalent width and length of the SIW cavity resonator is approximately determined by [59]

\[ w_{eff} = w - \frac{d^2}{0.95 p} \tag{2.32} \]

\[ l_{eff} = l - \frac{d^2}{0.95 p} \tag{2.33} \]

where \( p/d \) is smaller than 3 and \( d/w \) is smaller than 1/5. A more accurate empirical equation to extract the equivalent dimension is given as [60]

\[ x_{eff} = x - 1.08 \times \frac{d^2}{p} + 0.1 \times \frac{d^2}{x} (x = w, l) \tag{2.34} \]
Consequently, the fundamental resonant frequency $f_{\text{TE101}}$ of SIW cavity shown in Fig. 2.6 could be calculated using the following formula:

$$f_{\text{TE101}} = \frac{c_0}{2\pi\sqrt{\varepsilon_r}} \sqrt{\frac{1}{w_{\text{eff}}^2} + \frac{1}{l_{\text{eff}}^2}}$$  \hspace{1cm} (2.35)

2.3.4 Acoustic Resonator

As we know, the modern mobile communication systems are supporting multi-band and multi-standard operations. The adjacent channels are quite close to each other for efficient frequency spectrum usage. The traditional approach of adopting more resonators in filter design to improve channel selectivity will in turn result in higher insertion loss along the transmission path. Therefore, acoustic wave resonator, that makes use of piezoelectric effect in crystals, provides a promising solution to tackle that problem [61]. There are basically two categories of acoustic resonators: surface acoustic
wave (SAW) resonator and bulk acoustic wave (BAW) resonator, exhibited in Fig. 2.7. In SAW resonator, the surface acoustic wave is confined by interdigital fingers and its resonant frequency depends on the pitch width and the spacing between interdigital pitches. Due to limitation of achievable pitch spacing imposed by photolithography, SAW resonators are primarily used for developing filters working below 1 GHz. Unlike SAW resonator, BAW resonator generates vertically acoustic wave and standing acoustic wave is existing between top and bottom electrode. The frequency at which resonance occurs is determined by the thickness of piezoelectric substrate and the mass of electrodes. The detailed resonance conditions involving the piezoelectric material properties are beyond the focus of this dissertation.

![Illustration of (a) SAW resonator and (b) BAW resonator, respectively.](image)

Figure 2.7 Illustration of (a) SAW resonator and (b) BAW resonator, respectively.

2. 4 Generation Mechanisms of Transmission Zeros

It is mentioned earlier that a huge variety of electric systems are heavily exploring the limited resource of frequency spectrum, which has to be divided with care to accommodate more systems. RF signals belong to each system should be confined within the assigned frequency range by employing filters with high frequency selectivity. The
introduction of finite transmission zeros is an effective approach to enhance the sharpness of transition from passband to stopband. Due to the advantages of microstrip technology in terms of fabrication cost and integration with other active circuit elements, microstrip resonators are still widely used in modern wireless communication systems. Next, we will select microstrip resonator as demonstration example to illustrate several principles of introducing transmission zeros for improving frequency selectivity in filter design.

2.4.1 Cross coupling

In Fig. 2.8, a 3-order inline cascaded inline coupling topology for filter design is presented. Circle $S$, and $L$ is used to represent for input and output, respectively and the solid circles are basic functional resonators. The whole structure is arranged to be reciprocal and its $(N + 2) \times (N + 2)$ coupling matrix is written by,

$$
[m] = \begin{bmatrix}
0 & m_{s1} & 0 & 0 & 0 \\
0 & m_{s1} & m_{11} & m_{12} & 0 \\
0 & 0 & m_{12} & m_{22} & m_{23} \\
0 & 0 & 0 & m_{23} & m_{33} \\
0 & 0 & 0 & 0 & m_{3L} \\
\end{bmatrix}
$$

Based on [62], the coupling matrix is optimized to meet the following specifications:

- Center frequency $900$ MHz
- Bandwidth of passband $40$ MHz
- Return loss in the passband $<-25dB$
- Out-of-band rejection $>15$ dB for frequencies $\leq 850$ MHz and $\geq 950$ MHz

Fig. 2.9 plots the synthesized S-Parameters corresponding to the optimized coupling matrix given as
\[ [m] = \begin{bmatrix} 0 & 1.2 & 0 & 0 & 0 \\ 1.2 & 0 & 1.1 & 0 & 0 \\ 0 & 1.1 & 0 & 1.1 & 0 \\ 0 & 0 & 1.1 & 0 & 1.2 \\ 0 & 0 & 0 & 1.2 & 0 \end{bmatrix} \] 

(2.37)

Figure 2.8 A 3-order inline cascaded coupling topology.

\[ \text{Figure 2.9 Synthesized S-Parameters associated with the coupling matrix in Eq. 2.37.} \]

Obviously, no transmission zeros are observed from Fig. 2.9 and the reason could be explained by Eq 2.19, from which, the location of the transmission zero is determined by letting:

\[ \text{Num}\left([A]_{5,1}^{-1}\right) = m_{s1}m_{12}m_{23}m_{3L} = 0 \] 

(2.38)
Num means the numerator of the term in brackets. To satisfy above condition, either $m_{s1}/m_{3L}$ or $m_{12}/m_{23}$ is supposed be zero, which is not impossible. Consequently, there is no transmission zero close to the passband.

In Fig. 2.10, a revised 3-order coupling scheme with cross-coupling is introduced. The solid line denotes the direct coupling while the dashed line represents for the cross-coupling coupling. Therefore, from the source to the load, RF signals could pass through two transmission paths. Similarly, the coupling matrix is written by

$$
[m] = \begin{bmatrix}
0 & m_{s1} & 0 & 0 & 0 \\
m_{s1} & m_{11} & m_{12} & m_{13} & 0 \\
0 & m_{12} & m_{22} & m_{23} & 0 \\
0 & m_{13} & m_{23} & m_{33} & m_{3L} \\
0 & 0 & 0 & m_{3L} & 0
\end{bmatrix}
$$

(2.39)
When $m_{13}$ has non-zero value, the condition for the existence of transmission zero is:

$$Num([A]_{5,1}^{-1}) = m_{51}[m_{12}m_{23}m_{3L} - m_{13}m_{3L}(\Omega + m_{22})] = 0 \quad (2.40)$$

Obviously, the normalized position of transmission zero after some simple mathematical operations is obtained by,

$$\Omega = \frac{m_{12}m_{23}}{m_{13} - m_{22}} \quad (2.41)$$

If $m_{12}m_{23}/m_{13} > m_{22}$, the transmission zero will be appear on the right side of passband. If $m_{12}m_{23}/m_{13} < m_{22}$, there will be one transmission zero locating at the left side of passband. With above knowledge in mind as well as the specifications of filter such as,

- Center frequency $900 MHz$
- Bandwidth of passband $40 MHz$
- Return loss in the passband $<-25 dB$
- Out-of-band rejection $>20 dB$ for frequencies $\geq 950 MHz$

We apply the method in [62] so as to extract the optimized coupling matrix and it is achieved as

$$[m] = \begin{bmatrix} 0 & 1.25 & 0 & 0 & 0 \\ 1.25 & -0.28 & 1.08 & 0 & 0 \\ 0 & 1.08 & 0.52 & 1.08 & 0 \\ 0 & 0 & 1.08 & -0.28 & 1.25 \\ 0 & 0 & 0 & 1.25 & 0 \end{bmatrix} \quad (2.42)$$
Fig. 2.11 plots the synthesized results based on Eq. 2.42. Having obtained the coupling coefficients and external quality factor for the bandpass filter with desired performance, the detailed physical dimensions can be found by ANSYS HFSS to meet the requirements based on [55]. In this demonstration example, we select microstrip open loop resonators as the basic building blocks and the filter configuration is shown in Fig. 2.12. The simulated S-Parameters obtained by HFSS are shown in Fig. 2.13 and one

Figure 2.12 Demonstration example of a microstrip filter adopting 3-order coupling topology with cross-coupling (Substrate info: $\delta = 10.8$, $h = 1.27$ mm).

Figure 2.13 Simulated S-Parameters of the 3-order filter presented in Fig. 2.12.
transmission zero is found to appear at around 0.95 GHz.

2.4.2 Source-load coupling

In addition to introducing extra cross coupling between resonators to enable multi-path transmission between source and load, another effective approach of producing transmission zeros is to introduce direct source-load coupling. We take a 2-order coupling scheme as an instance and it is demonstrated in Fig. 2.14. In a 2-order coupling topology, the transmission zero generation condition is derived as,

\[
\text{Num}(A^{-1}) = 0 \tag{2.43}
\]

By expanding Num([A]_{4,1}^{-1}), we obtain,

\[
m_{SL}\Omega^2 + 2m_{SL}m_{11}\Omega + m_{SL}m_{11}^2 - m_{12}^2m_{SL} + m_{S1}m_{12} = 0 \tag{2.44}
\]

Figure 2.14 A 2-order coupling topology with source-load coupling for filter design.

Figure 2.15 The comparison of synthesized S-Parameters with/without source-load coupling.
From Eq 2.44, we notice that, if \( m_{SL} = 0 \), Eq. 2.44 does not hold anymore. No transmission zero will thus be generated around the passband. However, if \( m_{SL} \neq 0 \), and also \( \Delta = 4m_{SL}m_{12}(m_{SL}m_{12} - m_{S1}^2) > 0 \), there will be two different solutions existing for Eq. 2.44, resulting in two finite transmission zeros close to the passband. Fig. 2.15 plots a comparison between the synthesized S-Parameters with and without source-loading coupling. The center frequency and bandwidth of the demonstrated 2-order filter is set as 890 MHz and 40 MHz, respectively. To validate the theoretical analysis, a filter consisting of open loop resonators with source-load coupling is modeled in HFSS. Its detailed configuration and simulated S-Parameters using HFSS is given in Fig. 2.16 and Fig. 2.17. Two transmission zeros appear approximately at 0.8 GHz and 1.0 GHz.

Figure 2.16 Demonstration example of a microstrip filter adopting 2-order coupling topology with source-load coupling (Substrate info: \( \delta = 10.8 \), \( h = 1.27 \) mm).

Figure 2.17 Simulated S-Parameters of the 2-order filter presented in Fig. 2.16.
2.4.3 Skew-symmetrical feed

It is reported in [63] that different tapped-line feed points on resonators could also significantly influence the frequency response of filters. As shown in Fig. 2.18, Feed lines are tapped at different positions to excite two open loop resonators. It is indicated that there are two parallel transmission paths for the propagation of RF signals. Depending on the phase difference along two transmission paths, they are referred to as non-0° feed structure and 0° feed structure, respectively.

The coupling gaps between resonators are modeled as π–networks as provided in Fig. 2.19. Regarding the 2-order filter incorporating symmetrical input/output in Fig. 2.18 (a), it has been proven impossible to satisfy the condition for the existence of transmission zeros close to the passband if there is no source-loading coupling. However,
adopting the skew-symmetrical feed structure in Fig. 2.18 (b) will not only exhibit the same passband response but also create two extra transmission zeros close to and on the opposites of the passband. This 0° feed structure is analyzed using transmission matrices and the positions of transmission zeros are simply determined by

\[ \tan(\theta_1) + \tan(\theta_2) = \frac{1}{Z_0 \omega C_2} \]  

(2.45)

Since \( C_2 \) is small, it means that the transmission zeros will be at the frequencies when \( \theta_1 \approx \pi/2 \) or \( \theta_2 \approx \pi/2 \). Thus,

\[ f_{Z1} = \frac{c_0}{4l_1 \sqrt{\varepsilon_{\text{eff}}}} \]  

(2.46)

\[ f_{Z2} = \frac{c_0}{4l_2 \sqrt{\varepsilon_{\text{eff}}}} \]  

(2.47)

To illustrate the effect of adopting 0° feed structure, we simulate two 2-order filters with non-0° feed structure and 0° feed structure in Fig. 2.20, and the simulated S-Parameters are plotted in Fig. 2.21. From the comparison, the selectivity and stopband rejection of the 2-order filter with 0° feed structure are significantly increased.

Figure 2.20 The detailed configuration of 2-order open loop filter incorporating symmetrical and skew-symmetrical feed structures (Substrate info: \( \delta = 10.8 \); \( h = 1.27 \) mm).
2.5 Conclusions

Generally, filters are crucial components in various electric systems that perform the functions of selecting or confining RF signals within certain frequency range. Over the past century, filter design method has evolved from traditional image-parameter theory, insertion loss theory to coupling matrix theory which is particularly suitable for designing coupled-resonator filters with elliptic or quasi-elliptic response. Based on specific application demands, a variety of RF resonators are proposed and they could be applied as the basic building blocks in the coupling matrix theory. To improve the frequency selectivity of coupled-resonator filters, finite transmission zeros could be generated with the introduction of cross-coupling, source-load coupling and the adoption of 0° feed structure, which are all illustrated in the following chapters.

Figure 2.21 Simulated S-Parameters of two 2-order open loop filters adopting symmetrical and skew-symmetrical feed structures.
CHAPTER 3
COMPACT DUAL-BAND FILTER USING OPEN/SHORT STUB LOADED STEPPED IMPEDANCE RESONATORS

3.1 Introduction

Dual-band filters with high performance, low cost and compact size are essential and highly demanded in current and future dual-band wireless communication systems. During the past decades, great industrial demands have significantly pushed forward the research in the field of dual-band bandpass filter [64]-[70]. Although stepped impedance resonators (SIRs) are used to precisely adjust the ratio of two specified passband center frequencies by changing the structural parameters [64], it is difficult to selectively control the bandwidth of individual passband. In [65]-[67], dual-band filters featuring controllable bandwidths are realized respectively on dual-, tri-, and quad-mode stub-loaded resonators. Better frequency selectivity is achieved for the dual-band filters in [68]-[70], however, the filter in [68] is based on multilayer technology which demands a lamination process, inevitably increasing fabrication difficulty and cost. In [69] and [70], the dual-band filters suffer from relatively narrow bandwidth of rejection band and complex resonator configuration, respectively. In general, it still remains a challenging task to realize controllable bandwidths, high frequency selectivity, and wide rejection band simultaneously in a single dual-band filter.
To solve the technical difficulties, a microstrip dual-band filter composed of OSLSIRs and SSLSIRs is proposed in this chapter. Two fourth-order cross coupling schemes are generated along different transmission paths, which will produce quasi-elliptic response at each passband and make its bandwidth independently controllable. In addition, a 0° feed structure is adopted to improve the roll-off skirt [63]. By properly adjusting the structure parameters of the OSLSIRs/SSLSIRs, their harmonic resonances are misaligned without affecting the in-band performance. After extensive analyses and parametric studies, a dual-band filter was designed, fabricated and measured.

3.2 Microstrip Bandpass Filter with Single-Band Response

Fig. 3.1 illustrates a novel fourth-order coupling scheme for a bandpass filter with single-band response and its physical implementation using microstrip technology. Based on Eq 2.19, we could obtain the normalized positions of filter transmission zeros by considering $S_{21} = 0$, and solving the following equations:

$$m_{14} \cdot \Omega^2 + [m_{13}^2 - m_{12}^2 + m_{14} \cdot (m_{22} + m_{33})] \cdot \Omega$$

$$+(m_{14} \cdot m_{22} - m_{12}^2) \cdot m_{33} + m_{13}^2 m_{22} = 0$$

(3.1)
When $m_{13}^2 - m_{12}^2 + m_{14} \cdot (m_{22} + m_{33}) = 0$, two normalized transmission zeros could be symmetrically located close to the passband. According to the desired specification of a single-band filter: center frequency $f_0 = 2.1$ GHz, bandwidth $BW = 290$ MHz, and the positions of two transmission zeros $f_{z1} = 1.75$ GHz, $f_{z2} = 2.45$ GHz. The coupling matrix $[m]$ at the working frequency is optimized by [62], as given in Eq. 3.2. Fig. 3.2 plots the synthesized S-Parameters corresponding to above coupling matrix.

$$
[m] = \begin{bmatrix}
0 & 1.3 & 0 & 0 & 0 & 0 \\
1.3 & 0.046 & 0.89 & 0.824 & -0.27 & 0 \\
0 & 0.89 & 1 & 0 & 0.89 & 0 \\
0 & 0.824 & 0 & -1 & -0.824 & 0 \\
0 & -0.27 & 0.89 & -0.824 & 0.046 & 1.3 \\
0 & 0 & 0 & -0.824 & 0 & 1.3
\end{bmatrix}
$$

(3.2)

In Fig. 3.1 (b), the T-stub loaded dual-mode resonator consisting of a $\lambda/2$ microstrip line and a T-shaped-stub attached at its center is embedded between two outer coupled half-wavelength resonators. Due to the symmetry of the T-stub loaded dual-mode resonator presented in Fig. 3.3, odd- and even-mode theory is applied and there will be perfect electric conductor wall and perfect magnetic conductor wall existing along the

![Figure 3.2 Synthesized S-Parameters corresponding to Eq. 3.2.](image_url)
symmetrical line of the dual-mode resonator. For the purpose of simplicity, \( w_b \) is assumed equal to \( 2w_a \). Based on the transmission line theory, the input admittance looking into the odd-mode and even mode equivalent half circuit are respectively obtained by

\[
Y_{\text{in, odd}} = \frac{1}{jZ_0 \tan \left( \frac{\beta l_a}{2} \right)}
\]

\[
Y_{\text{in, even}} = \frac{1}{-jZ_0 \cot \left[ \beta \left( \frac{l_a}{2} + l_b \right) \right]}
\]

where \( Z_0 \) is the characteristic impedance of microstrip line whose width is \( w_a \). By setting \( \text{Im}(Y_{\text{in, odd}}) = 0 \) and \( \text{Im}(Y_{\text{in, even}}) = 0 \), the resonance occurs when

\[
f_{\text{odd}} = \frac{c}{2l_a \sqrt{\varepsilon_{\text{eff}}}}
\]

Figure 3.3 Illustration of T-stub loaded dual-mode resonator and its odd-/even-mode equivalent half circuit.

Figure 3.4 Schematic layout of the finally optimized fourth-order filter with multiple transmission zeros and wide rejection band.
With known coupling coefficients and external quality factor, the coupling gaps between

\[ f_{\text{even}} = \frac{c}{(l_a+2l_b)\sqrt{\varepsilon_{\text{eff}}}} \]  

(3.6)

resonators and the input/output tapped position are uniquely determined by [55].

In order to further improve the frequency response of developed single-band filter, two modifications are made to the geometrical layout, as illustrated in Fig. 3.4. Firstly, the symmetrical input/output is replaced by skew-symmetrical which is also named 0° feed structure. As clearly demonstrated in Chapter 1, the 0° feed structure will produce

![Diagram](image)

Figure 3.5 (a) The layout of the lowpass structure and (b) the schematic of short circuit, anti-coupled line connected by an open-circuited line.

![Graph](image)

Figure 3.6 (a) Simulated S-parameters of the lowpass structure and (b) simulated S21 of the fourth-order filter with and without the lowpass structure, respectively.
two more transmission zero around the passband. Furthermore, a pair of anti-coupled short-circuited at one end by capacitive load section is inserted at the input and output feed, which can add several transmission zeros to widen the stop-band bandwidth. Fig. 3.5 shows the lowpass structure including two anti-coupled high-impedance lines and a low impedance line. The cutoff frequency of the lowpass structure is far beyond the passband of the original fourth-order filter and its nearby transmission zeros. Fig. 3.6 presents the simulated S-parameters of the lowpass structure and the simulated $S_{21}$ of the fourth-order filter with and without the lowpass structure. Obviously, excellent out-of-band harmonic suppression is achieved by using the lowpass structure.

For the purpose of verification, the proposed single-band filter is fabricated on single-layer printed circuit board (PCB) of 1mm in thickness, and its relative permittivity is $\varepsilon_r = 2.65$. Optimized by ANSYS HFSS, the geometrical sizes are summarized as follows: $L_1 = 2.9 \text{ mm}$, $L_2 = 18.6 \text{ mm}$, $L_3 = 12.2 \text{ mm}$, $L_4 = 3.3 \text{ mm}$, $L_5 = 3.1 \text{ mm}$, $L_6 = 10.55 \text{ mm}$, $L_7 = 15 \text{ mm}$, $L_8 = 2.65 \text{ mm}$, $L_9 = 13.4 \text{ mm}$, $L_{10} = 4.0 \text{ mm}$, $L_{11} =$

Figure 3.7 Simulated and measured S-parameters of the single-band filter in (a) narrowband and (b) wideband, respectively.
6.5 mm, $L_{12} = 3.2$ mm, $L_a = 5.2$ mm, $L_b = 5.0$ mm, $W_1 = 0.8$ mm, $W_2 = 0.5$ mm, $W_3 = 1.0$ mm, $W_a = 0.3$ mm, $W_b = 4.0$ mm, $g_1 = 0.82$ mm, $g_2 = 1.3$ mm, $g_a = 0.3$ mm. And the input and output 50 Ω microstrip lines are set to be 2.7 mm, respectively. The simulated and measured S-parameters of the proposed filter are plotted in Figs. 3.7 (a) and (b). The central frequency $f_0$ is at 2.13 GHz, with its 3-dB bandwidth of 285 MHz. The measured insertion loss of the proposed filter is around 1.3 dB. Four transmission zeros are located at 1.79 GHz, 1.92 GHz, 2.26 GHz and 2.76 GHz, respectively. The existence of four transmission zeros contributes to high frequency selectivity. In addition, the 20-dB rejection band bandwidth is extended to $4.9f_0$.

![Detail of the SSLSIR 2 via](image)

Figure 3.8 Detailed configuration of the proposed dual-band BPF.

3.3 Microstrip Bandpass Filter with Dual-Band Response

Inheriting the same four-order cross coupling topology from the previous single-band filter, Fig. 3.8 illustrates the circuit configuration of the proposed dual-band BPF which is designed to operate at $f_1 = 1.63$ GHz, $f_2 = 2.73$ GHz with FBW $\Delta_1 = 8\%$ and $\Delta_2 = 5.5\%$. The dual-band BPF comprises of two identical OSLSIRs and two SSLSIRs of different sizes. SSLSIR 1 is coupled with fundamental odd-mode of OSLSIRs to yield
lower passband at $f_1$ and SSLSIR 2 is coupled with fundamental even-mode of OSLSIRs to constitute upper passband at $f_2$. This special arrangement is aiming to make the proposed dual-band BPF compact in size and realize a novel fourth-order cross coupling scheme at respective band, as illustrated in Fig. 3.9. Fig. 3.10 plots the simulated S-parameters of two cases in the inset of the figure and the proposed dual-band filter prototype for comparison. On the basis of the coupling scheme, quasi-elliptic responses are generated at the lower and upper passband. Besides, the in-band performance of lower/upper passband is not affected while SSLSIR2/SSLSIR1 does not exist.
The schematic layouts of OSLSIR and SSLSIR are given in Fig. 3.11 (a) and (b), respectively. For simplicity, we assume $Z_1 = 2Z_3$, $Z_4 = 2Z_6$ and define $R = Z_1/Z_2$, $R' = Z_4/Z_5$. The resonance conditions derived by odd-even-mode method are formulated as

$$\tan \theta_1 \cdot \tan \theta_2 = \frac{1}{R} \quad (3.7)$$

$$\frac{\tan \theta_2}{\tan(\theta_1 + \theta_3)} = \frac{1}{R} \quad (3.8)$$

$$\tan \theta_4 \cdot \tan \theta_5 = \frac{1}{R'} \quad (3.9)$$

$$\tan(\theta_4 + \theta_6) \cdot \tan \theta_5 = \frac{1}{R'} \quad (3.10)$$

where Eq. 3.7, 3.8 and 3.9, 3.10 are applied to determine the resonant frequencies of the OSLSIR and SSLSIR, respectively. OSLSIR is taken for an example to derive the relationship between fundamental modes $f_{odd,0}$, $f_{even,0}$ and harmonic modes $f_{odd,1}$, $f_{even,1}$. According to Eq. 3.7 and 3.8, Fig. 3.12 depicts the first harmonic mode $f_{odd,1}$ and $f_{even,1}$ normalized by the fundamental mode $f_{odd,0}$ and $f_{even,0}$ under various combinations of impedance ratios and length ratios. The fundamental odd- and even-mode resonant frequencies $f_{odd,0}$, $f_{even,0}$ of the OSLSIR are utilized as the two passband center frequencies, i.e., $f_1$ and $f_2$. And once the design specifications of passband center frequencies $f_1$ and $f_2$ are given, we are able to locate the harmonic modes $f_{odd,1}$, $f_{even,1}$ by the following procedures:

1). Select proper impedance ratio $R$ and length ratio $\alpha$ which equals to $\theta_1/(\theta_1 + \theta_2)$, the value of $\theta_1$ and $\theta_2$ could be obtained by Eq. 3.7;

2). Calculate the physical length $l_1$, $l_2$ with $\theta_1$, $f_1$ by

$$l_1 = \frac{c\theta_1}{2\pi f_1 \sqrt{\varepsilon_{eff}}} \quad (2.31)$$

$$l_2 = (1/\alpha - 1) \cdot l_1 \quad (2.32)$$
3). Based on Eq. 3.8 and the known value of $R, l_1, l_2$ and $f_2, \theta_3$ can be simply obtained and $\gamma$ is determined by $(\theta_1 + \theta_3)/(\theta_1 + \theta_2 + \theta_3)$;

![Diagram](image)

Figure 3.11 Schematic layout of (a) the OSLSIR and (b) the SSLSIR utilized in this design.

![Graph](image)

Figure 3.12 The first odd-mode harmonic $f_{odd,1}$ normalized by $f_{odd,0}$ and the first even-mode harmonic $f_{even,1}$ normalized by $f_{even,0}$.

4). Refer to Fig. 3.12, locate the frequency ratios corresponding to $\alpha, \gamma, and R$.

It is presented in Fig. 3.13 (a) that the surface current distribution of the dual-band filter at two passband frequencies. The signals are delivered through different transmission paths while the dual-band BPF is operating at $f_1$ and $f_2$, making the bandwidth of each passband independently controllable. As a consequence, the
bandwidth requirements could be easily fulfilled. As exhibited in Fig. 3.13 (b), the bandwidth of the lower passband increases if $g_2$ decreases. Meanwhile, $g_2$ has no impact

Figure 3.13 (a) Electric current distribution on the surface of the dual-band filter and (b) simulated bandwidths of each band under different values of $g_2$ and $g_3$. 
on the bandwidth of the upper passband. It becomes opposite when $g_3$ is adjusted.

Fig. 3.14 demonstrates that four transmission zeros TZ2, TZ3, TZ5, TZ6 generated by the fourth-order cross coupling scheme are controlled by the gap $g_1$ between two outer OSLSIRs. At the locations of three extra transmission zeros TZ1, TZ4, TZ7, the length from the tap position denoted by $d_4$ to three different ends of OSLSIR is quarter wavelength, thus shorting RF signals to ground as designed. It is concluded that the transmission zeros obtained by two different generation principles can be manipulated.

For the purpose of demonstration, a dual-band BPF is designed and fabricated by using a single-layered printed circuit board (PCB) process on Rogers 4350B with the permittivity of $3.48 \pm 0.05$, the loss tangent of 0.003 and the thickness of 0.5mm. The layout dimensions adopted for the dual-band filter are obtained by ANSYS HFSS and summarized as follows (unit: mm): $L_1 = 15.6$, $L_2 = 7.4$, $L_3 = 9.05$, $L_4 = 14.3$, $L_5 =
$L_6 = 1.4, \ L_7 = 8.26, \ L_8 = 4.6, \ L_9 = 7.4, \ W_1 = 1.0, \ W_2 = 2.8, \ W_3 = 2.0,$
$W_4 = 1.0, \ W_5 = 2.85, \ W_6 = 1.6, \ W_7 = 1.0, \ W_8 = 2.0, \ W_9 = 2.0, \ d_1 = 4.74, \ d_2 = 1.68, \ d_3 = 0.6, \ g_1 = 0.52, \ g_2 = 0.66, \ g_3 = 0.67$ with $R$ and $\alpha$ finally specified as 2.1 and 0.68. The diameters of two via-holes in SSLSIRs are both set to be 0.5 mm.

Figure 3.15 Simulated and measured S-parameters of the dual-band filter in (a) narrowband and (b) wideband, respectively.
The simulated and measured S-parameters of the fabricated dual-band filter sample in narrowband and wideband are respectively plotted in Fig. 3.15 (a) and (b) as well as its photograph in the inset of Fig. 3.15 (b). The measured center frequency and 3-dB bandwidth of the first passband are 1.58 GHz and 118 MHz. The second passband has the center frequency of 2.66 GHz and 3-dB bandwidth of 135 MHz. The discrepancies are caused by the variation of substrate’s permittivity and fabrication tolerance. The minimum insertion losses at two passbands are measured to be 1.5 and 2.15 dB including the extra losses from the SMA connectors. The 20-dB rejection band has been extended to 8.5 GHz, which is benefited from the distribution of the first four resonant modes of the OSLSIR and two distinct SSLSIRs as illustrated in Fig. 3.16. The spurious band due to the strongly excited even-mode third harmonic of OSLSIR is occurring at around 8.5 GHz [71], which may be suppressed by adding extra defected ground structure (DGS) to further extend the rejection band. Finally, Table 3.1 compares the results of the proposed dual-band filter with those of other similar works. As noticed, our approach introduces less insertion losses and one addition transmission zero compared with [70], but at the expense of increasing 30% area.

Figure 3.16 The resonant modes of each resonator appearing below 9 GHz.
In this section, a dual-band BPF using open/short stub loaded stepped impedance resonators is proposed and implemented. Both the center frequencies and the bandwidths of each passband of the dual-band filter could be individually controlled. By taking advantage of a novel fourth-order cross coupling scheme and skew-symmetrical feeding structure, high frequency selectivity is achieved. Furthermore, the second harmonics of OSLSIRs and SSLSIRs are purposely misaligned to extend the rejection band. With the features of compactness, controllable bandwidths, high skirt selectivity and wide rejection band obtained, the proposed dual-band filter is possibly suitable for dual-band communication systems in the future.

<table>
<thead>
<tr>
<th>Table 3.1</th>
<th>Comparison With Related Reference (TZS Denotes Transmission Zeros and UR Denotes -20 dB Rejection Band)</th>
</tr>
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<tr>
<td>Ref.</td>
<td>IL(dB)</td>
</tr>
<tr>
<td>[67]</td>
<td>1.11/0.4</td>
</tr>
<tr>
<td>[68]</td>
<td>0.4/0.58</td>
</tr>
<tr>
<td>[69]</td>
<td>2.1/2.2</td>
</tr>
<tr>
<td>[70]</td>
<td>2.65/2.44</td>
</tr>
<tr>
<td>Proposed</td>
<td>1.5/2.15</td>
</tr>
</tbody>
</table>

3.4 Conclusion

This chapter mainly demonstrates the design of a compact dual-band filter with high frequency selectivity, wide rejection band, and individually controllable bandwidth. Firstly, the fourth-order cross coupling topology is analyzed first and then applied to implement a single-band filter. Skew-symmetrical feed (0° feed) is utilized to introduce two more transmission zeros close to the passband for sharp roll-off skirt. In the single-band filter design, an intuitive way of inserting lowpass structures is employed to suppress the harmonics, thus extending the rejection band. Afterwards, the same fourth-
order cross coupling scheme is applied in the dual-band filter design. At respective band, RF signals are propagating through different paths, making the bandwidth of each passband individually controllable. The occupied dual-mode resonators in the dual-band filter design are all made into stepped impedance form instead of uniform impedance form. As a result, the harmonic peaks are all staggered, and rejection band is extended.
CHAPTER 4

BANDPASS FILTERING POWER Divider WITH SHARP ROLL-OFF SKIRT AND ENHANCED IN-BAND ISOLATION

4.1 Introduction

Recent development of RF/microwave circuits requires high level of integration, miniaturization and multi-function. Power dividers and filters are indispensable components in RF front-end of transceiver. And they both occupy considerable circuit area especially in low Giga-Hertz range. To solve this issue, integrating two or more separate circuit topologies has become a popular concept [72]-[75]. Although the fourth-order bandpass power dividers in [72] and [22] possess good frequency selectivity, the in-band isolation deteriorates accompany with the high-order response. In [73], the second harmonic is suppressed by finite transmission zeros contributed by a novel third-order coupling topology. However, the process of drilling and metallizing holes inevitably increases the fabrication complexities and cost. The power dividers with filtering response in [74] and [75] exhibit good isolation performance, but the operation bandwidth is relatively narrow [74] and the frequency selectivity requires to be further improved [75].

A compact filtering power divider with multiple transmission zeros around the passband and enhanced in-band isolation is reported in this section. It is composed of
four dual-mode resonators which are chosen for reducing circuit size and realizing an inter-cross-coupled coupling topology. The generation principle of transmission zeros is fully analyzed and the mixed electric and magnetic couplings between coupled dual-mode resonators will induce extra transmission zeros near the passband. To obtain a good isolation performance, two isolation resistors of different values are placed at the center of dual-mode resonators between channels. Based on the proposed idea, dual functions of bandpass response exhibiting sharp roll-off skirt and power division with enhanced in-band isolation are simultaneously achieved.

Figure 4.1 Detailed layout of the proposed filtering power divider.

Figure 4.2 Coupling scheme of the filtering power divider.
4.2 Circuit Design

Fig. 4.1 illustrates the physical layout of the proposed filtering power divider with operation frequency $f_0$ of 4.45 GHz and fractional bandwidth FBW of 7%. It could be simply treated that two identical filters consisting of two coupled dual-mode resonators are connected by a T-junction at input port. Open stub loaded dual-mode resonators are chosen here to reduce the total number of resonators needed for realizing a fourth-order bandpass response. Odd- and even- mode frequencies of each dual-mode resonator are adjusted to be the center frequency of the bandpass PD. It is worth pointing out that the special arrangement of two coupled dual-mode resonators as shown in Fig. 4.1 is able to replace the $\lambda/4$-line based on the transformation technology proposed in [21]. Fig. 4.2 presents the corresponding inter-cross-coupled coupling diagram of proposed PD. Either odd- or even- mode of one dual-mode resonator is coupled to odd- and even- modes of another resonator at the same time. All open stubs are bent at the end to gain extra freedoms in accommodating the even-mode frequencies and coupling coefficients between two even-modes.

If even-mode excitation is applied to output ports, the topology of the equivalent

![Figure 4.3 Coupling matrix and its determined transmission response.](image-url)

Figure 4.3 Coupling matrix and its determined transmission response.
bandpass filtering function is shown in the inset plot of Fig. 4.3 and this filter is designed to have the same center frequency and bandwidth as the targeted PD. Fig. 4.3 shows the synthesized transmission response as well as \((n+2)\)-order coupling matrix at center frequency \(f_0\) [56]. The locations of the transmission zeros could be found by solving

\[ S_{21} = -2j[A]^{-1}_{o,1} = 0 \]  

(4.1)

where \([A]\) is a function of coupling matrix [56], and therefore we could obtain

\[ \Omega^2 = \left( m_{s2}^2 m_{13} m_{24} + m_{s1}^2 m_{24} m_{13} + m_{s1} m_{s2} m_{23} m_{14} \right) \left( m_{s1} m_{s2} (m_{23} + m_{14}) \right) \]  

(4.2)

where \(\Omega = \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right) / FBW\) is the normalized angular frequency.

Figure 4.4 Parameter \(\Omega^2\) expressed in (4.2) and the normalized angular frequency square as a function of frequency.
It is also known from [76] that the coupling between two odd-modes in our design are actually mixed electric and magnetic coupling, which will make the coupling coefficient \( m_{13} \) in Fig. 3 become frequency-variant. Thus, \( m_{13} \) is expressed by

\[
m_{13}(\omega) = m_{13}(\omega_0) \cdot \omega(\omega_0^2 - \omega_m^2)/[\omega_0(\omega^2 - \omega_m^2)]
\]

(4.3)

Where \( \omega_m = 2\pi f_m = (L_m C_m)^{-\frac{1}{2}} \) and \( L_m, C_m \) respectively denotes the coupling inductance and capacitance. Obviously, it is difficult to directly get the analytical solution of (2.39) once \( m_{13} \) becomes frequency-variant, so we plot \( \Omega^2 \) in (4.2) and normalized

Figure 4.5 Synthesized \( S_{21} \) for different values of \( f_m \) when mixed coupling is implemented between (a) mode 1 and 3, and (b) mode 2 and 4, respectively.
angular frequency square simultaneously in Fig. 4.4 to graphically locate the transmission zeros. As shown in Fig. 4.7, there are only two finite transmission zeros produced around the passband while frequency-variant mixed coupling does not exist between two odd-modes, which has been illustrated in Fig. 4.3. It is easily observed that an extra transmission zero will be generated above the passband on the premise of $f_m > f_0$ when mixed coupling is introduced. And as expected, its position depends on $f_m$. For validation, Fig. 4.5 (a) plots the synthesized S-parameters for different $f_m$ while $m_{13}$ is a frequency-dependent variable expressed by Eq. 4.3. Besides, the open-stubs of dual-mode resonators are all bent at the end to create mixed coupling between two even-modes of adjacent dual-mode resonators. This configuration will lead to larger coupling capacitance between even-modes than that between odd-modes because maximum electric fields concentrate at the open gaps [76]. Similarly, by replacing the fixed coupling coefficient $m_{24}(\omega_0)$ in Fig. 4.3 with a frequency-dependent $m_{24}(\omega)$ as in Eq. 4.3, an extra transmission zero will appear below the passband under the condition that $f_m < f_0$. Fig. 4.5 (b) depicts the synthesized S-parameters for various values of $f_m$ when $m_{24}$ becomes a frequency-variant coupling coefficient.

When mixed coupling is existing between mode 1, 3 and between mode 2, 4 concurrently, it is anticipated that two extra transmission zeros will be produced apart from two transmission zeros close to the passband as shown in Fig. 4.3. The anticipation is further confirmed by analyzing the proposed filtering PD in ANSYS HFSS, with its transmission response $S_{21}$ ($S_{31}$) given in Fig. 4.6. The mixed coupling between odd-modes 1 and 3 contributes to TZ4 while the mixed coupling between even-modes 2 and 4 accounts for the generation of TZ1. In addition, the bandwidth control mechanism of
proposed device is also included in Fig. 4.6. The bandwidth shrinks as the coupling gap denoted by \( g \) increases.

4.3 Implementation and Discussion

Based on previous theoretical analyses and parametric studies, a power divider with BPF response has been implemented to operate at 4.45 GHz with fractional bandwidth of 7%. It is fabricated on a single-layer PCB with the relative permittivity of \( \varepsilon_r = 2.65 \), loss tangent \( \tan(\delta) = 0.003 \), and the thickness \( h = 1 \text{ mm} \). The geometrical dimensions of the filtering PD are determined as follows (Unit: mm): \( l_1 = 3, l_2 = 3, l_3 = 12.4, l_4 = 0.5, l_5 = 12.6, l_6 = 1.1, l_7 = 3, l_8 = 3, l_9 = 5.5, l_{10} = 4.5, w_1 = 2.67, w_2 = 1.5, w_3 = 0.2, w_4 = 0.4, w_5 = 1.6, w_6 = 1.5, w_7 = 1.5, w_8 = 2.67, d = 9.3, d = 0.5 \). Using the method in [73], the adopted resistance values \( R_1 \) and \( R_2 \) of two isolation resistors are decided to be 50 and 300 ohm, respectively, so as to achieve good matching status at output ports.

![Graph](image.png)

Figure 4.6 Simulated \( S_{21}(S_{31}) \) for different values of \( g \) as a function of frequency.
Finally, the S-parameters of the filtering PD is measured using Keysight 8722ES Vector Network Analyzer. Fig. 4.7 plots the simulated and measured performance, with good agreement obtained between them. The acceptable discrepancy between simulated and measured $S_{11}$ is mainly attributed to that the hand-soldering positions of two isolation resistors could not be precisely controlled. As observed, the fabricated filtering power divider is functioning at 4.48 GHz, with absolute 3-dB bandwidth of 310 MHz. The measured insertion losses of two channels are around (3+1.3) dB and (3+1.35) dB. It is worth mentioning that the isolation performance within whole passband is better than 18 dB. Four finite transmission zeros are positioned at 3.85 GHz, 4.25 GHz, 4.75 GHz and 5.4 GHz, respectively, resulting in good frequency selectivity.

![Figure 4.7 Simulated and measured S-parameters of the proposed bandpass filtering PD (insert). (Dotted line: simulated results; Solid line: measured results)](image)

4.4 Conclusion

This chapter presents a compact device demonstrating dual functions of frequency selection and power division. A novel inter-cross-coupled coupling scheme containing
mixed coupling is fully investigated and utilized to produce multiple transmission zeros around the passband. Two surface-mounted resistors are placed between four dual-mode resonators, which are useful to provide enhanced isolation between output ports. The feasibility of the proposed design methodology is verified by the agreement between theoretical and experimental data of implemented filtering PD. The proposed device has potential application for future wireless communication systems.
CHAPTER 5

BALANCED DUAL-BAND FILTER WITH FIXED/RECONFIGURABLE CENTER FREQUENCIES

5.1 Introduction

Balanced circuit architectures have attracted tremendous attention from both academia and industry in design of modern communication system, especially under the trend of system-on-chip (SoC) or system-in-package (SiP). With the requirements of high-density integration and miniaturization, interference and crosstalk, which highly degrade the performance of conventional single-ended unbalanced circuits, have become a severer problem than ever. RF and microwave filters are the most essential and critical components in the RF front ends of both the receiver and transmitter. Differential filters are known to have high immunity to environmental noise and electromagnetic interference as well as good dynamic range. Therefore, increasing attention and great effort has been devoted to the development of high performance microwave differential circuits [77], [78].

Balanced filters, as crucial components in a balanced T/R system, have already been investigated for a long time. Research efforts are devoted to develop single-band response with wide stopband [79] and common-mode suppression [80]. With the ever-increasing demand of dual-band and multi-band microwave communication systems,
balanced dual-band filters are becoming key components in the systems because they could greatly reduce the circuit size and insertion loss by replacing the single-ended dual-band filter in combination with baluns (balanced-to-unbalanced). However, there is relatively little research reported on the balanced dual-band filters [80]-[84]. In [80], quarter-wavelength and half-wavelength resonators are used as the building blocks under differential-mode and common-mode excitation, respectively. In [81]-[84], different types of stepped-impedance resonators are utilized to realize dual-band response. And the balanced dual-band filter based on substrate integrated waveguide presented recently [85] could only achieve two closely separated passbands. Although stub-loaded multi-mode resonator has been utilized quite often to design multi-band single-ended filters [86], to the authors’ best knowledge, no application has been found that uses the open stub loaded short-ended dual-mode resonator in a balanced dual-band second-order filter design.

5.2 Balanced Dual-Band Filter with Fixed Center Frequencies

In this section, a balanced dual-band filter with high DM frequency selectivity and enhanced CM suppression is presented. The center frequency of each passband under

Figure 5.1 Physical layout of the proposed balanced dual-band filter.
DM excitation is determined by the even- and odd-mode frequency of the stub-loaded resonator, respectively. By loading a stepped-impedance stub, good CM suppression in a broad range could be achieved. Source-load coupling introduces transmission zeros which will not only improve the frequency selectivity in DM response but also enhance the CM suppression level within DM passbands. A prototype balanced filter is designed and fabricated in this section.

Fig. 5.1 illustrates the detailed physical layout of the proposed balanced dual-band filter. Two identical DOSLRs are referred to as resonator A and B for convenience. A stepped-impedance stub is attached to the center of resonator A. The whole structure is arranged symmetric with respect to the dashed line to prevent conversion between DM and CM signals. In that regard, a perfect electric conductor wall and a perfect magnetic conductor wall will be present along the symmetric line of the DOSLR under DM and CM excitation, respectively, as shown in Fig. 5.2.

Figure 5.2 (a) A dual-open-stub-loaded resonator (DOSLR) and its equivalent half structure under (b) DM and (c) CM excitation.
By properly arranging the position of two open-stubs, the DM equivalent half structure could work as a short-ended resonator loaded by an open-stub at the center as illustrated in Fig. 5.2 (b). Therefore, odd- and even-mode analysis is adopted to analyze its resonant frequencies, which is expressed by

$$f_{\text{odd}} = \frac{c}{2L_a \sqrt{\varepsilon_{\text{eff}}}}$$  \hspace{1cm} (5.1)

where $c$ is the light speed in free space and $\varepsilon_{\text{eff}}$ denotes the effective dielectric constant of the substrate. The even-mode input impedance seen from the open end without considering the effect of discontinuities at the connection point is expressed as

$$Z_{\text{in, even}} = j \frac{2Y_b \tan(\beta L_a) + 4Y_a \tan(\beta L_b)}{2Y_a Y_b - Y_b^2 \tan(\beta L_a) \tan(\beta L_b)}$$  \hspace{1cm} (5.2)

According to resonance condition which is $\text{Im}(Y_{\text{in, even}}) = 0$, the resonance occurs when

$$\tan(\beta L_a) \tan(\beta L_b) = \frac{2Y_a}{Y_b}$$  \hspace{1cm} (5.3)

In the special case of $Y_b = 2Y_a$,

$$f_{\text{even}} = \frac{c}{4(L_a + L_b) \sqrt{\varepsilon_{\text{eff}}}}$$  \hspace{1cm} (5.4)

It is evident that $2f_{\text{even}} < f_{\text{odd}}$, which is consistent with our design objective where $\frac{f_2}{f_1} = \frac{5.2}{2.4} > 2$. Under common-mode operation, the input admittance for the half structure as shown in Fig. 5.2 (c) seen from the end of transmission line is given by

$$Y_{\text{in,CM}} = j \frac{Y_a^2 \tan^2(\beta L_a) + Y_a Y_b \tan(\beta L_b) \tan(\beta L_b) - Y_a^2}{2Y_a \tan(\beta L_a) - Y_b \tan^2(\beta L_a) \tan(\beta L_b)}$$  \hspace{1cm} (5.5)

When $\text{Im}(Y_{\text{in, CM}}) = 0$, the resonant condition can be obtained as follows

$$\tan(2\beta L_a) \tan(\beta L_b) = \frac{2Y_a}{Y_b}$$  \hspace{1cm} (5.6)

Again, in the special case of $Y_b = 2Y_a$. 

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\[ f_{CM} = \frac{(2n-1)c}{4(2L_a+L_b)\sqrt{\varepsilon_{eff}}} \quad n = 1,2,3,\ldots \quad (5.7) \]

$L_b$ is selected to be smaller than $L_a$ in our design to meet the design objective, which results in the fact that the CM second harmonic will appear between two desired DM passbands due to the following relationship:

\[ \frac{c}{4(L_a+L_b)\sqrt{\varepsilon_{eff}}} < \frac{3c}{4(2L_a+L_b)\sqrt{\varepsilon_{eff}}} < \frac{c}{2L_a\sqrt{\varepsilon_{eff}}} \quad (5.8) \]

As shown in Fig. 5.3 (a), under the special condition of $Y_b = 2Y_a$, the DM even-

![Figure 5.3](image1)

(a)

![Figure 5.3](image2)

(b)

Figure 5.3 (a) Resonant frequency of the DOSLR under DM and CM operation versus $L_b$ with different values of $L_a$. (b) The simulated DM and CM responses of the initial design and schematic layout (inserted).
mode frequency could be conveniently tuned to the desired position by adjusting the length $L_b$ of the open-stub while the odd-mode frequency is determined by $L_a$ which is fixed. When $L_b$ is smaller than $L_a/2$, only the second harmonic of CM response is located between two DM passbands. When $L_b$ becomes larger than $L_a/2$, there will be an intercepting point for the DM odd mode frequency and the CM third harmonic. The results in Fig. 5.3 (a) are verified with the simulated DM and CM responses of the initial design as shown in Fig. 5.3 (b). The CM second harmonic leads to an unsatisfied CM suppression level of 10 dB around 3.75 GHz. In order to obtain a better CM suppression level of 20 dB in a relatively large range, covering from lower DM passband to upper DM passband, a stub is attached to the center of Resonator A to shift the second harmonic of resonator A away from that of resonator B. It is noted from Fig. 5.1 that the stub is in the form of stepped-impedance instead of the uniform-impedance for miniaturization.

Compared with other balanced dual-band filters in [82]-[84], better frequency selectivity in DM operation could be achieved by the finite transmission zeros around each passband, generated by the source-load coupling. It is worth mentioning that the controllable finite transmission zeros will not only improve frequency selectivity of DM response, but also enhance the suppression level of CM response within two desired DM passbands. Fig. 5.4 (a) and (b) presents the influence of stepped-impedance stub and source-load coupling on the CM response without deteriorating the DM performance. The second harmonic of resonator A controlled by $l_8$ and (or) $w_5$ in CM response is misaligned with that of resonator B. Under different combinations of $l_8$ and $w_5$, the DM response keeps unaltered as shown in Fig. 5.4 (a) because the stepped-impedance stub is
short-circuited under DM excitation. Besides, the size of the stepped-impedance stub will also affect the position of TZ2. From Fig. 5.4 (b), the variance of source-load coupling strength will only shift TZ1, without changing the position of TZ2 and the CM second harmonic of resonator A. Additionally, it is under our expectation that the DM

![Image](https://via.placeholder.com/150)

(a)

![Image](https://via.placeholder.com/150)

(b)

Figure 5.4 The simulated DM and CM transmission responses versus (a) \(l_9\) and \(w_s\), and (b) \(l_{11}\), respectively.
transmission zeros marked by three black solid circles will move towards passbands with the increase of source-load coupling strength.

To verify the proposed idea, a balanced dual-band filter sample operating at 2.4 and 5.2 GHz is designed, which is implemented on a single-layer PCB with the relative permittivity of $\varepsilon_r = 2.65$, $\tan \delta = 2.65$, and the thickness $h = 1\text{mm}$. Eight via-holes with the diameter of $d_{\text{via}} = 0.5\text{mm}$ are drilled and metallized. The geometrical sizes of the balanced dual-band filter optimized by HFSS are summarized as follows (Unit: mm):

$l_1 = 30.8$, $l_2 = 21$, $l_3 = 11.35$, $l_4 = 3.15$, $l_5 = 2.5$, $l_6 = 3$, $l_7 = 13.75$, $l_8 = 2.5$, $l_9 = 3.5$, $l_{10} = 2.35$, $l_{11} = 1.1$, $l = 12.25$ $w_1 = 1$, $w_2 = 1$, $w_3 = 0.85$, $w_4 = 1$, $w_5 = 4$, $w_6 = 0.5$, $w = 2.65$, $d_1 = 2.4$, $d_2 = 0.6$, $d_3 = 0.55$. Fig. 5.5 depicts the measured CM and DM responses as well as the simulated results for comparison. The fabricated sample are centering at 2.49/5.35 GHz. The small frequency discrepancies between measurement and simulations are due to the unpredictable fabrication tolerance. The measured

![Figure 5.5](image)

Figure 5.5 Simulated (dashed lines) and measured (solid lines) responses of the proposed balanced dual-band filter.
insertion losses at two DM passbands are around 2.64 dB and 2.58 dB, which are slightly larger than the simulated results as they include the extra losses from the SMA connectors used for measurement. Three measured DM transmission zeros are found at 2.75, 5.0 and 5.5 GHz, respectively. The measured CM suppression level of 20 dB is observed from the lower to the upper DM passband. Specially, the CM suppression level is up to 50 dB within two DM passbands. Table 5.1 summarizes the comparison of the insertion loss, CM attenuation, number of DM transmission zeros, and circuit size with the state of art technologies. The balanced dual-band filter in this work has achieved significantly enhanced CM suppression level and high DM frequency selectivity simultaneously.

In this section, a novel balanced dual-band filter based on dual-open-stub-loaded resonators (DOSLRs) has been implemented. The resonant frequencies of the DOSLR under differential-mode and common-mode excitation have been fully analyzed. High differential-mode frequency selectivity and enhanced common-mode suppression are observed. The measured results are also in good agreement with the simulated ones, which validates the feasibility of the presented idea.

<table>
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<th>Ref.</th>
<th>$f_0$ (GHz)</th>
<th>IL(dB)</th>
<th>Atten. (dB)</th>
<th>TZs</th>
<th>Size($\lambda g^2$)</th>
</tr>
</thead>
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<td>[17]</td>
<td>2.44/5.57</td>
<td>1.78/2.53</td>
<td>30/27</td>
<td>1</td>
<td>0.33×0.31</td>
</tr>
<tr>
<td>[18]</td>
<td>2.44/5.25</td>
<td>2.4/2.82</td>
<td>53/47</td>
<td>0</td>
<td>0.5×0.46</td>
</tr>
<tr>
<td>[21]</td>
<td>2.4/3.57</td>
<td>0.87/1.9</td>
<td>28/31</td>
<td>1</td>
<td>0.5×0.2</td>
</tr>
<tr>
<td><strong>This Work</strong></td>
<td><strong>2.49/5.35</strong></td>
<td><strong>2.64/2.58</strong></td>
<td><strong>51/50</strong></td>
<td><strong>3</strong></td>
<td><strong>1×0.14</strong></td>
</tr>
</tbody>
</table>

5.3 Balanced Dual-Band Filter with Tunable Center Frequencies

Due to the ever-increasing demand of multi-standard operation, some single-band balanced BPFs in [87], [88] are configured to provide tunable center frequencies with
constant absolute/fractional bandwidth (ABW/FBW). It is also noticed that the design in [89] achieves a wide tuning range of both center frequency and bandwidth. However, to the best knowledge of authors, there are no balanced dual-band filters with tunable center frequencies reported so far [90]. In this section, a varactor-incorporated doubly short-ended resonator is firstly introduced and its resonant frequencies are extensively studied under differential-mode and common-mode excitation, respectively. In the design, the inter-stage coupling coefficient for each differential-mode passband is flexibly controlled, and source-load coupling generates quasi-elliptic DM passband response and produces common-mode transmission zeros. A stepped impedance stub is loaded at the symmetric plane to separate CM resonances for further suppression of common-mode resonance. Finally, a balanced dual-band filter with tunable center frequencies is designed, fabricated, and measured to demonstrate the efficacy of the proposed design concept.

Figure 5.6 presents circuit configuration of the proposed balanced dual-band BPF with reconfigurable center frequencies. It employs two doubly short-ended resonators
loaded with two short-ended stubs. Varactor diodes are loaded as shown in Fig. 5.7 (a) to enable frequency tuning capability. A voltage null is known to appear at the center point O under differential-mode excitation. As a matter of fact, two red sine curves in Fig. 5.7 (a) define the DM voltage distribution at \( f_{DM,1} \) and \( f_{DM,2} \) without consideration of two short-ended stubs circled in green. After loading two stubs at point A and B, where the voltage of DM signal characterized by \( f_{DM,2} \) is null, the associated DM resonant frequency \( f_{DM,2} \) keeps constant while the other DM resonant frequency \( f_{DM,1} \) is changed.

It is worth mentioning that a back-to-back varactor pair is inserted at the center to ensure that two loaded stubs are consistently located at voltage null of DM signal \( f_{DM,2} \) under different voltage bias conditions of varactor diodes. According to Fig. 5.7 (b), the DM input admittances are expressed as

\[
Y_{in,odd} = j\omega C_1 - jY_a cot(\beta L_a) \tag{5.9}
\]

\[
Y_{in,even} = \frac{j\omega C_2}{2} + \frac{Y_b}{2} \cdot \frac{2Y_{int} + jY_b tan(\beta L_b)}{Y_b + j2Y_{int} tan(\beta L_b)} \tag{5.10}
\]

\[
Y_{int} = Y_a \cdot \frac{\omega C_1 + jY_a tan(\beta L_a)}{Y_a - \omega C_1 tan(\beta L_a)} \tag{5.11}
\]

Figure 5.7 (a) Varactor diode-incorporated doubly short-ended resonator with two short-ended stubs and (b) its odd-mode and even-mode equivalent half-circuit under DM excitation.
By applying $\text{Im} \left( Y_{\text{in,odd}}(\omega_{\text{odd}}) \right) = 0$ and $\text{Im} \left( Y_{\text{in,even}}(\omega_{\text{even}}) \right) = 0$ where $\omega_{\text{odd}} = 2\pi f_{\text{odd}}$, and $\omega_{\text{even}} = 2\pi f_{\text{even}}$. Thus, two DM resonant frequencies are obtained and plotted in Fig. 5.8 to illustrate their relationship with varactor values $C_1$ and $C_2$, respectively. Similarly, the CM resonant frequency is determined by solving,

$$
\frac{\omega C_1 + Y_a \tan(\beta L_a)}{Y_a - \omega C_1 \tan(\beta L_a)} + \frac{\omega C_2 + Y_b \tan(\beta L_b)}{Y_b - \omega C_2 \tan(\beta L_b)} = Y_a \cot(\beta L_a)
$$

(5.12)

A stepped-impedance stub is loaded at the center to ensure that CM resonances of two

![Graph showing DM resonant frequencies against varactor values](image)

Figure 5.8 DM resonant frequencies $f_{DM,1}$ and $f_{DM,2}$ against (a) $C_1 \ (C_2 = 1.8 \ pF)$ and (b) $C_2 \ (C_1 = 1.8 \ pF)$, respectively ($Y_a = Y_b = 0.0115\,\text{S}$, and $L_b = 7.5\,\text{mm}$ on single-layer PCB with $\varepsilon_r = 2.65$, thickness $h = 1\,\text{mm}$).

![Diagram of DM bi-section structure](image)

Figure 5.9 DM bi-section structure and its corresponding coupling scheme.
resonators are separated enough and no CM peak appears within the tuning range of DM passbands [80].

Figure 5.10 Simplified schematic layout of the coupled structures constituting two DM passbands at (a) $f_{DM,2}$ and (b) $f_{DM,1}$.

Under differential-mode operation, the bi-section structure functions as a conventional dual-band filter with two-pole response. Its coupling scheme in Fig. 5.9 illustrates that there are two transmission paths for DM signal $f_{DM,1}$, while only one path is existing for DM signal $f_{DM,2}$. Source-load coupling indicated by dashed line will generate DM transmission zeros to improve frequency selectivity and CM transmission zeros to enhance suppression level within DM passbands. As a demonstration example, lower passband is designed at 3.2 GHz with a fractional bandwidth of 5 % and the center frequency of higher passband is selected at 4.35 GHz with a fractional bandwidth of 4.2 %. Based on above specifications, the coupling matrices at $f_{DM,1}$ and $f_{DM,2}$ are respectively obtained as: $M_{S1}^L = M_{L2}^L = 1.17$, $M_{12}^L = 1.75$, $M_{SL}^L = -0.1$ and $M_{S1}^H = M_{L2}^H = 1.18$, $M_{12}^H = 1.72$, $M_{SL}^H = -0.12$. Furthermore, the inter-stage coupling coefficient $K_{12}$ and external quality factor $Q_e$ are calculated as: $K_{12}^L = 0.105$, $Q_e^L = 12.2$, and $K_{12}^H = 0.086$, $Q_e^H = 14.4$.

To examine the relationship between coupling gap and inter-stage coupling coefficient, simplified layouts of two coupled structures constructing DM passbands are
shown in Fig. 5.10 and then investigated. Fig. 5.10 (a) stands for the coupled structures at higher DM passband. With given value of coupling length \( l_7 \) and capacitance \( C_1 \), the coupling coefficient \( K_{12}^H \) uniquely depends on \( d_1 \). In Fig. 5.10 (b), extra coupling

![Graph showing coupling coefficient \( K_{12}^H \) versus frequency for different \( d_1 \) values.](image)

**Figure 5.11** Coupling coefficient \( K_{12}^H \) versus \( d_1 \) (\( l_7 = 3.5\, \text{mm}, \, C_1 = 5\, \text{pF} \)) and (b) coupling coefficient \( K_{12}^L \) versus \( d_4 \) (while \( d_1 \) is fixed at 1.5 mm (\( l_5 = 4.5\, \text{mm}, \, C_1 = C_2 = 5\, \text{pF} \)).
controlled by $d_4$, $l_5$, $C_2$ is introduced at lower DM passband and it represents the
coupling between two loaded stubs. According to [88], coupling coefficient $K_{12}^H$ and $K_{12}^L$
versus different gap sizes are plotted in Fig. 5.11, which can be referred to meet the
requirement of each bandwidth. In addition, the external quality factors with respect to
d_2 and l_3 are extracted and plotted in Fig. 5.12, based on $Q_e = f_0 / \Delta f_{\pm 90^\circ}$.

![Graph](image1)

(a)

![Graph](image2)

(b)

Figure 5.12 External quality factors as a function of (a) $d_2$ when $l_3 = 15.5mm$, and
(b) $l_3$ when $d_2 = 0.3mm$ ($C_1 = C_2 = 5pF$).
In order to validate the presented design methodology, a balanced dual-band BPF with tunable center frequencies is designed, simulated and fabricated on a single-layered printed circuit board (PCB) substrate with a dielectric constant of 2.65, loss tangent of 0.003 and thickness of 1 mm. Varactor diodes are hyperabrupt junction SMV 1763 ($R_s \sim 0.5\Omega$, $C_j \sim 1.8 - 9.0pF$ at $V_c \sim 4.5 - 0V$) from Skyworks Solutions Inc. The 3-D structure is firstly simulated in ANSYS HFSS with tuning elements considered as ideal capacitors. The initial capacitance values for $C_1$ and $C_2$ are 5 pF and 1.8 pF. The simulated structure is then imported to Keysight ADS and varactor diode’s spice model from manufacturer is employed to accomplish the simulation of frequency tuning. Final layout dimensions are optimized and summarized as follows (unit: mm): $l_1 = 33$, $l_2 = 13.5$, $l_3 = 15.5$, $l_4 = 3.7$, $l_5 = 4.5$, $l_6 = 5.6$, $l_7 = 5$, $l_8 = 7$, $l_9 = 1.3$, $l_{10} = 3.3$, $l_{11} = 4$, $l_{12} = 4$, $w_1 = 1$, $w_2 = 1$, $w_3 = 0.85$, $w_4 = 0.3$, $w_5 = 1$, $w_6 = 4$, $d_1 = 1.5$, $d_2 = 0.23$, $d_3 = 0.35$, $d_4 = 3.3$, $d_s = 1$, $d_{via} = 0.5$. Four surface mounted capacitors with capacitance value of $C_0 = 47pF$ are adopted to isolate different DC bias voltages $V_1$ and $V_2$.

Figure 5.13 Comparison between simulation and measurement results as $V_1$ and $V_2$ are set to 4.5 V.
Additionally, eight surface mounted inductors with inductance value of $L_0 = 100\, nH$ are attached between the soldering pads and the resonators for RF choke.

The simulated and measured S-parameters are shown in Fig. 5.13 as $V_1$ and $V_2$ are both set to 4.5 V. Fig. 5.14 (a) shows the measured results as $V_1$ varies from 1, 2, 3, 4, to 4.5 V while $V_2$ is fixed at 4.5 V. As observed, two DM passbands are tuned toward higher frequencies simultaneously. Fig. 5.14 (b) presents the measured results when $V_2$ is chosen as 1, 2, 3, 4, 4.5 V and $V_1$ remains a constant value of 4.5 V. Obviously, only the lower DM passband shifts upwards and the higher passband keeps unchanged. With no doubt, it is also feasible to adjust the higher passband frequency and fix the lower one by tuning $V_1$ and $V_2$ together. The frequency shift of lower passband caused by altering $V_1$ can be easily compensated by adjusting the value of $V_2$. At last, Fig. 5.15 plots the variations of each passband insertion loss and 3-dB bandwidth corresponding to the tuning process shown in Fig. 5.14.

Figure 5.14 Measured results of tunable balanced dual-band BPF when only (a) $V_1$ and (b) $V_2$ is tuned.
5.4 Conclusion

In this part, balanced dual-band filters with fixed/reconfigurable center frequencies based stub-loaded are demonstrated and implemented. They are designed using dual-stub-loaded resonators. The differential-mode and common-mode resonant frequencies are fully analyzed. A stepped impedance stub is loaded to separate CM resonances of two resonators to prevent CM peaks. Two coupling regions are created to flexibly control the bandwidths of DM passbands. By changing the applied voltages on diodes, the passband frequencies are tunable with good passband performances maintained. As a conclusion, the proposed design has wide applications in a reconfigurable dual-band front-end system with high electromagnetic interference and environmental noise.

![Figure 5.15 Variations of insertion loss and 3-dB bandwidth when (a) V₁ and (b) V₂ is adjusted, respectively.](image)
CHAPTER 6
EQUATION-BASED SOLUTIONS TO MICROSTRIP TABBED ROUTINGS FOR CROSSTALK REDUCTION

6.1 Introduction

Over the last few decades, there has been a progressively expanded tendency of increasing trace density on printed circuit boards (PCBs), driven by the requirement of low-cost and highly compact wireless communication systems, such as miniaturized Multiple-Input Multiple-Output (MIMO) systems [91]. Within a limited space, signal traces at a high data transfer rate along with other closely placed active traces are vulnerably subject to electric and magnetic coupling, which is known as crosstalk. Nowadays, crosstalk has already become one of the dominant limiting factors for achieving higher data transfer rate. To reduce far-end crosstalk (FEXT), various approaches have been explored by researchers so far [92]-[96]. Surface tab-routing is recently proposed by Intel Corporation for crosstalk mitigation to improve double data rate (DDR) channel performance [97].

Fig. 6.1 clearly illustrates the comparison results of far-end crosstalk under three different scenarios for DDR routing traces. When two microstrip traces of 50 Ω are apart from each other with enough spacing, there is little crosstalk issue existing (Example (1), Fig. 6.1). Nevertheless, the far-end crosstalk degrades severely the received signal as the same traces are getting closer in highly compact printed circuit boards (Example (2), Fig.
In order to overcome the obstacle and meet the demands of high speed and high-density integration simultaneously, interdigital trapezoidal tabs are introduced between two closely coupled microstrip traces, which exhibits lower far end crosstalk (Examples (3), Fig. 6.1). However, how to optimize the tabbed routing structures efficiently and accurately remains a question. To the authors’ knowledge, there are no analytical models with satisfactory accuracy to address the complex structure preferably in the form of S-parameters, except for the time-consuming numerical modeling approach.

With the above-mentioned problem in mind, generalized equation-based solutions need to be developed considering the coupled, asymmetrical, lossy, and non-uniform microstrip structures. In [98], four-port terminal characteristic parameters for asymmetrical coupled lines in an inhomogeneous system were analyzed, which introduced the $c$-mode and $\pi$-mode in principle for the first time. In [99], there was a detailed discussion on the relationship between the geometrical dimensions of coupled striplines with unequal widths and their admittance matrix, but microstrip structures were

![Figure 6.1](image.png)

**Figure 6.1** Comparison of far-end crosstalk (FEXT) in three different configurations (Example (1) and (2) are only different in the spacing between traces; Example (2) and (3) have the same spacing but example (3) is distinguished from (2) by the introduction of interdigital trapezoidal tabs).
not discussed. Although in [100], various forms of asymmetrical coupled lines, such as coplanar strips, striplines, and microstrip lines, are analyzed in terms of capacitance matrices, the trace thickness is assumed to be zero, which is not true in reality.

In this chapter, we successfully develop a set of complete analytical equation-based solutions to predict the behaviors of tabbed routing microstrip lines in a coupled fashion, which has been used to mitigate far-end crosstalk. As an alternative approach to the lengthy numerical simulations, the proposed equation-based solutions are applicable not only to coupled uniform microstrip lines with equal/unequal trace widths but also to the transitional structures with linearly varying trace width, which can be treated as a cascade of many segmented sub-sections with different trace widths. Furthermore, a new concept to define and calculate the fringing capacitances between the asymmetrically coupled tab-sections, is introduced as Tab-Coupling Fringing Capacitance (TCFC), which is especially useful for improving accuracy in the process of extracting the far-end crosstalk. It is demonstrated that our equation-based solutions are targeted at practical high speed and high-density PCB designs with an optimization process in mind, where the common physical characteristics of transmission lines, such as inhomogeneous, coupled, asymmetrical, lossy, and non-uniform properties, are fully addressed.

The chapter is organized as follows: Section 6.2 introduces the capacitance and inductance matrices of asymmetrical coupled microstrip lines with unequal trace widths and inhomogeneous media. Section 6.3 discusses how to convert the lossless capacitance and inductance matrices into lossy and causal RLGC parameters. After that, S-parameters of asymmetrical coupled uniform microstrip traces with losses are derived. In section 6.4, the proposed methodology is extended to solve coupled, asymmetrical, lossy, and non-
uniform microstrip traces incorporating interdigital trapezoidal tabs. In each section, analytical results are compared with the counterparts obtained from commercial tools. Section 6.5 validates the efficacy of our proposed equation-based solutions in optimization of surface tab routing. In the end, we conclude our approaches and application aspects in section 6.6.

![Diagram](image)

Figure 6.2 (a) Cross-section view of coupled, asymmetrical microstrip lines denoted with their geometrical parameters and (b) illustration of basic capacitive components of various meanings in the configuration of coupled microstrip lines.

6.2 Expression for Capacitance and Inductance Matrices

As well known, conventional odd- and even- mode concepts defined in the case of identical coupled lines could not be directly applied when their widths become unequal. Propagation modes along asymmetrical coupled microstrip lines shown in Fig. 6.2 (a) are referred to as c- and π-mode [98], whose properties could be determined from associated capacitance and inductance matrices.

By taking the approach of dividing total capacitance into several capacitances pertaining to their physical nature, as illustrated in Fig. 6.2 (b), capacitance matrix $C$ of coupled microstrip structures is given as follow,

$$
C = \begin{bmatrix}
C_{11} + C_m & -C_m \\
-C_m & C_{22} + C_m
\end{bmatrix}
$$

(6.1)
Where \( C_{11} \) and \( C_{22} \) represent the capacitance of each line to ground and \( C_m \) is the mutual capacitance. They are formulated, respectively, by

\[
C_{11} = C_f + C_{p1} + C_{fe} \tag{6.2}
\]
\[
C_{22} = C_f + C_{p2} + C_{fe} \tag{6.3}
\]
\[
C_m = C_{m1} + C_{m2} + C_{m3} \tag{6.4}
\]

The analytical expressions of external fringing capacitance \( C_f \) and internal fringing capacitance \( C_{fe} \) are readily available by assuming trace-to-ground capacitive elements \( C_f \) and \( C_{fe} \) in the case of microstrip line being approximately half of those in the stripline configuration with substrate thickness of \((2b + t)\). From [99], \( C_f \) and \( C_{fe} \) are given, respectively, as

\[
\frac{C_f}{\varepsilon_r \varepsilon_0} = \frac{1}{\pi} \left[ 2t_b \ln(t_b + 1) - (t_b - 1) \ln(t_b^2 - 1) \right] \tag{6.5}
\]
\[
t_b = \frac{2b + t}{2b} \tag{6.6}
\]
\[
\frac{C_{fe}}{\varepsilon_r \varepsilon_0} = A_0 \left\{ \frac{s}{2b + t} - \frac{2}{\pi} \ln \left[ \cosh \left( \frac{\pi s}{2(2b + t)} \right) \right] \right\} + B_0 \tag{6.7}
\]

Where \( A_0 \) and \( B_0 \) are defined in the Appendix. \( C_{pi} \) \((i = 1, 2)\) and \( C_{m2} \) are simply determined by the following parallel-plate capacitor formulas,

\[
C_{pi} = \frac{\varepsilon_r \varepsilon_0 W_i}{b} \tag{6.8}
\]
\[
C_{m2} = \varepsilon_0 \frac{t}{s} \tag{6.9}
\]

The definition of gap capacitance in the air above microstrip traces \( C_{m1} \) is inherited from the case of coplanar strips without thickness in [100], and expressed as below

\[
C_{m1} = \varepsilon_0 \frac{\kappa(k')}{\kappa(k)} - \Delta C_f \tag{6.10}
\]
\[
k = \frac{1 + W_1/s + W_2/s}{(1 + W_1/s)(1 + W_2/s)} \tag{6.11}
\]
\[
k' = 1 - k^2 \tag{6.12}
\]

\( \Delta C_f \) is an empirical air-gap capacitance correction factor introduced in [100] and given in the Appendix. At last, gap capacitance in the dielectric \( C_{m3} \) is given by [100]
The capacitance matrix $C_{m3} = \sqrt{C_{d1}^* C_{d2}}$ \hspace{1cm} (6.13)

where $C_{di}$ ($i = 1, 2$) is described in the Appendix as well. In summation, the final capacitance matrix $C$ of asymmetrical coupled microstrip lines under the quasi-static condition is constructed directly from their cross-sectional physical parameters (Eq. 6.1). It is worth pointing out that the constructed capacitance matrix $C$ has taken the practical thickness of the trace into consideration, which was neglected in previous work \[100\]. It is known that in the homogeneous media, inductance matrix $L$ of coupled transmission lines is related to their corresponding capacitance matrix by

$L = \begin{bmatrix} L_{11} & L_m \\ L_m & L_{22} \end{bmatrix} = \frac{1}{v_s^2} \cdot U_{2 \times 2} \cdot C^{-1}$ \hspace{1cm} (6.14)

where $v_s$ is the wave velocity propagating in the homogeneous media surrounding the traces and $U_{2 \times 2}$ is a $2 \times 2$ identity matrix. However, due to the phenomenon of propagation mode velocity splitting in inhomogeneous media \[100\], \[101\], the inductance matrix of coupled microstrip lines could not be directly calculated using (6.14) from its own capacitance matrix $C$. The solution we adopt here is to replace the dielectric material with air, making the two coupled lines surrounded by homogeneous media air, and then generate a new dummy capacitance matrix $C_{\text{air}}$, which could be plugged back into (6.14).

<table>
<thead>
<tr>
<th>$C_{11}$ (pF)</th>
<th>Analytical</th>
<th>Q2D</th>
<th>ERROR (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>50</td>
<td>49</td>
<td>2</td>
<td></td>
</tr>
<tr>
<td>$C_{22}$ (pF)</td>
<td>66.7</td>
<td>68.6</td>
<td>2.8</td>
</tr>
<tr>
<td>$C_m$ (pF)</td>
<td>11.7</td>
<td>11.4</td>
<td>2.6</td>
</tr>
<tr>
<td>$L_{11}$ (nH)</td>
<td>490.8</td>
<td>477.5</td>
<td>2.7</td>
</tr>
<tr>
<td>$L_{22}$ (nH)</td>
<td>393.1</td>
<td>381.8</td>
<td>2.9</td>
</tr>
<tr>
<td>$L_m$ (nH)</td>
<td>117.3</td>
<td>112.6</td>
<td>4</td>
</tr>
</tbody>
</table>

**TABLE 6.1**

**Comparison of Capacitance Matrix and Inductance Matrix Obtained from Analytical and Numerical Approaches**
to obtain the inductance matrix needed [102]. There is a comparison of capacitance and inductance matrices between derived analytical approach and ANSYS Q2D in Table 6.1, demonstrating a good agreement between them. The structural parameters of arbitrarily selected microstrip line structures under analysis are chosen as: \( W_1 = 102 \, \mu m, \, W_2 = 201 \, \mu m, \, s = 165 \, \mu m, \, t = 30.48 \, \mu m, \, b = 200 \, \mu m. \)

6.3 Asymmetrical Coupled Microstrip Traces with Unequal Widths

The quasi-static capacitance and inductance matrices for asymmetrical coupled microstrip lines are investigated in detail and derived directly from their cross-sectional geometrical dimensions in previous section. Fig. 6.3 exhibits the extracted equivalent circuit model of asymmetrical coupled microstrip lines without considering any losses. While taking the DC resistive loss, skin effect contributed RF loss along with dielectric loss into consideration, frequency dependent causal RLGC parameters are formulated as follows [101], [103], [104]

![Figure 6.3 Equivalent circuit models of coupled lossless microstrip lines with unequal widths.](image-url)
\[
\hat{R} = \begin{bmatrix}
R_{dc,1} + R_{dc,1}G_p\sqrt{f/f_{surf}} & 0 \\
0 & R_{dc,2} + R_{dc,2}G_p\sqrt{f/f_{surf}}
\end{bmatrix}
\]  \quad (6.15-a)

\[
\hat{L} = \begin{bmatrix}
L_{11} + \frac{R_{dc,1}G_p}{2\pi\sqrt{f/f_{surf}}} & L_m \\
L_m & L_{22} + \frac{R_{dc,2}G_p}{2\pi\sqrt{f/f_{surf}}}
\end{bmatrix}
\]  \quad (6.15-b)

\[
\hat{G} = \begin{bmatrix}
\omega K(C_{11} + C_{m3}) & -\omega K C_{m3} \\
-\omega K C_{m3} & \omega K(C_{22} + C_{m3})
\end{bmatrix}
\]  \quad (6.15-c)

\[
\hat{C} = \begin{bmatrix}
P(C_{11} + C_{m3}) + C_{m1} + C_{m2} & -PC_{m3} - C_{m1} - C_{m2} \\
-PC_{m3} - C_{m1} - C_{m2} & P(C_{11} + C_{m3}) + C_{m1} + C_{m2}
\end{bmatrix}
\]  \quad (6.15-d)

where \(R_{dc,i} = \rho/(W_it) \quad (i = 1, 2)\) stands for the DC resistance per unit length of the \(i\)th trace whose width and thickness are defined by \(W_i\) and \(t\), respectively. Besides, \(f_{surf} = 4\rho/(t^2\pi\mu_0)\) denotes the skin-effect onset frequency and \(G_p\) is introduced for the conductor loss contributed by the current in the ground plane [101]. In dielectric material, \(K\) and \(P\) are defined, respectively, as

\[
K = -\frac{\varepsilon_r''}{\varepsilon_r'} \quad (6.16)
\]

\[
P = \frac{\varepsilon_r'}{\varepsilon_r} \quad (6.17)
\]

where frequency-dependent parts of dielectric material \(\varepsilon_r'\) and \(\varepsilon_r''\) are written,

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>(\varepsilon_r)</td>
<td>3.65</td>
</tr>
<tr>
<td>(\varepsilon_0)</td>
<td>(8.854 \times 10^{-12}) F/m</td>
</tr>
<tr>
<td>(\mu_0)</td>
<td>(4\pi \times 10^{-7}) H/m</td>
</tr>
<tr>
<td>(A_{sub})</td>
<td>0.045</td>
</tr>
<tr>
<td>(\tau_1)</td>
<td>1.6 ps</td>
</tr>
<tr>
<td>(\tau_2)</td>
<td>1.6 ms</td>
</tr>
<tr>
<td>(G_p)</td>
<td>1.5</td>
</tr>
<tr>
<td>(\rho)</td>
<td>(1.764 \times 10^{-8}) (\Omega m)</td>
</tr>
</tbody>
</table>
respectively, by
\[
\begin{align*}
\varepsilon_\tau' &= \varepsilon_r + \frac{A_{su} \ln r}{2} \left[ \frac{\tau_2^2(1+\omega^2\tau_1^2)}{\tau_1^2(1+\omega^2\tau_2^2)} \right] \\
\varepsilon_\tau'' &= A_{su} [\arctan(\omega \tau_1) - \arctan(\omega \tau_2)]
\end{align*}
\] (6.18)

Table 6.2 lists all the parameters which are not particularly specified but are randomly chosen for the numerical simulation as a demonstration example. After that, the self-impedance \( z_i \) \((i = 1, 2)\) and admittance \( y_i \) \((i = 1, 2)\) per unit length of each trace as well as the mutual impedance \( z_m \) and admittance \( y_m \) per unit length are given by [11], [15]
\[
\begin{align*}
z_1 &= R_{dc,1} + R_{dc,1}G_p\sqrt{f/f_{surf}} + j\omega[L_{11} + R_{dc,1} G_p/(2\pi\sqrt{f/f_{surf}})] \\
z_2 &= R_{dc,2} + R_{dc,2}G_p\sqrt{f/f_{surf}} + j\omega[L_{22} + R_{dc,2} G_p/(2\pi\sqrt{f/f_{surf}})] \\
z_m &= j\omega L_m \\
y_1 &= \omega K(C_{11} + C_{m3}) + j\omega[P(C_{11} + C_{m3}) + C_{m1} + C_{m2}] \\
y_2 &= \omega K(C_{22} + C_{m3}) + j\omega[P(C_{22} + C_{m3}) + C_{m1} + C_{m2}] \\
y_m &= -\omega K C_{m3} + j\omega(-PC_{m3} - C_{m1} - C_{m2})
\end{align*}
\] (6.19-6.24)

Subsequently, referring to [98], \(a_1, a_2, b_1, b_2\) are defined for constructing the analytical expression of propagation constants \(\gamma_c\) and \(\gamma_\pi\) for in-phase and antiphase waves, along with the voltage ratios \(R_c/R_\pi\) between two traces for each mode. All of the explicit parameters are addressed in a causal and frequency-dependent fashion at this point.
\[
\begin{align*}
a_1 &= y_1 z_1 + y_m z_m \\
a_2 &= y_2 z_2 + y_m z_m \\
b_1 &= z_1 y_m + y_2 z_m \\
b_2 &= z_2 y_m + y_1 z_m \\
\gamma_c^2 &= \frac{a_1 + a_2}{2} + \frac{1}{2}[(a_1 - a_2)^2 + 4b_1 b_2]^2 \\
\gamma_\pi^2 &= \frac{a_1 + a_2}{2} - \frac{1}{2}[(a_1 - a_2)^2 + 4b_1 b_2]^2 \\
R_c &= \frac{1}{2b_1} \left\{ (a_2 - a_1) + [(a_2 - a_1)^2 + 4b_1 b_2]^{1/2} \right\} \\
R_\pi &= \frac{1}{2b_1} \left\{ (a_2 - a_1) - [(a_2 - a_1)^2 + 4b_1 b_2]^{1/2} \right\}
\end{align*}
\] (6.25-6.27)

Then, the characteristic impedance of two traces corresponding to \(c\)-mode and \(\pi\)-mode are generated by
\[
Z_{c1} = \frac{1}{\gamma_c} \frac{z_1 z_2 - z_m^2}{z_2 - z_m R_c} \\
\] (6.28-a)
\[ Z_{c2} = \frac{R_c z_1 z_2 - z_m^2}{\gamma_c z_1 R_c - z_m} \]  
(6.28-b)

\[ Z_{\pi 1} = \frac{1}{\gamma_{\pi}} \frac{z_1 z_2 - z_m^2}{z_1 R_{\pi} - z_m} \]  
(6.28-c)

\[ Z_{\pi 2} = \frac{R_{\pi} z_1 z_2 - z_m^2}{\gamma_{\pi} z_1 R_{\pi} - z_m} \]  
(6.28-d)

As indicated in [98], each element in a 4×4 Z-matrix of asymmetrical coupled transmission lines with length \( l \), surrounded by inhomogeneous media, is determined with the equation based representations of \( \gamma_c, \gamma_{\pi}, R_c, R_{\pi}, Z_{c1}, Z_{c2}, Z_{\pi 1}, Z_{\pi 2} \), and the conversion between Z-matrix and S-matrix is achieved by [106]

\[ [S] = ([Z]/Z_0 + [U_{4x4}])^{-1}([Z]/Z_0 - [U_{4x4}]) \]  
(6.29)

where \( U_{4x4} \) is a 4×4 identity matrix and \( Z_0 \) is the reference impedance.

To evaluate the accuracy of the derived equation-based solutions for obtaining the

![Graphs showing comparisons between analytical and numerical results for coupled asymmetrical microstrip lossy traces.](image)

Figure 6.4 Comparison between analytical and numerical results for coupled asymmetrical microstrip lossy traces whose geometrical dimensions are summarized as: 
\( W_1 = 300 \mu m, W_2 = 600 \mu m, s = 500 \mu m, t = 30.48 \mu m, b = 200 \mu m, l = 50 \text{ mm} \).
S-parameters of the coupled asymmetrical microstrip transmission lines, numerical simulation is performed in comparison with ANSYS HFSS and results are plotted in Fig. 6.4, together with the analytical results produced in this part for confirmation. It is observed that the analytical and numerical simulation results match with each other very well in Fig. 6.4, thus confirming the accuracy of proposed method in the form of the S-parameters, which are targeted at coupled, asymmetrical, and lossy microstrip traces with unequal widths, whose dimensions are summarized in the caption of Fig. 6.4.

In order to clearly demonstrate the causality of the proposed analytical closed-form solution, a well-defined time-domain pulse described in [103] is launched at one end of the narrow line and observed at the other end. Its length is increased to 20 cm and other dimensions are kept the same. The S4P file coming from our developed closed-form solution as well as that from HFSS with/without causality enforcement are imported into Keysight ADS to compare the starting points of received pulse, respectively. As shown in Fig. 6.5, the proposed RLGC(\(f\)) model is causal.

![Circuit diagram and time-domain waveforms comparison](image.png)

Figure 6.5 Circuit of the time-domain simulation performed in ADS and comparison of the time-domain waveforms at the receiving end.
6.4 Analysis of Microstrip Tab-Routing

As reported in [97], adding tabs between closely placed single-ended DDR traces is beneficial for reducing far-end crosstalk. This section is aimed to demonstrate the feasibility of adopting the proposed approach for presenting the scattering parameters for the surface tab-routing traces. Fig. 6.6 presents a detailed schematic layout of coupled, asymmetrical, lossy, and non-uniform microstrip traces with interdigital trapezoid tabs incorporated for crosstalk mitigation. Instead of performing time-consuming numerical simulations to obtain the scattering parameters, the proposed analytical method is demonstrated to apply the equivalent model shown in Fig. 6.7, to the optimization process with fast computation and accurate results. As noticed, the coupled microstrip lines with interdigital tabs consist of many coupled microstrip line sections. It turns out to be feasible by deriving the scattering parameters of each section first and cascading them thereafter for obtaining the total S-parameters.

Note that section Q1 and Q2 in red, as in Fig. 6.7 are asymmetrical coupled microstrip lines with unequal widths, which have been investigated earlier in section 6.3

Figure 6.6 Detailed geometrical dimensions of two coupled microstrip lines incorporating trapezoid tabs.
of this paper. The input/output sections and the middle section are coupled microstrip traces with equal width but different in length. In fact, their scattering parameters can be immediately known upon applying the proposed analytical approach as a special case with $W_1$ being equal to $W_2$. Next, analysis of transitional section is performed in Part 6.4.1 below and a new concept of tab coupling fringing capacitance (TCFC) is uniquely introduced in Part 6.4.2.

Figure 6.7 Equivalent model of asymmetrical coupled non-uniform microstrip lossy lines with interdigital trapezoid tabs.

Figure 6.8 Segmentation of the magnified transition structure T1 to apply proposed analytical method for its scattering parameters.
6.4.1 Transitional Section with Linearly Varying Trace Width

When it comes to the transitional section connecting high- and low-impedance trace for reducing reflections caused by impedance discontinuity, such as T1, T2, T3, and T4 in Fig. 6.7, directly applying the developed equation-based solutions may not be suitable noting that the slopes on each side of the tab section render unequal width coupling. However, thanks to the characteristic of linearly varying trace width, the method of segmentation can be employed and the T1 section in Fig. 6.7 is taken as an example to show the feasibility of applying the developed analytical approach to the transitional structure. Fig. 6.8 depicts the concept of segmentation and section T1 is segmented into \( m \) sub-sections. The physical dimensions \((w_0, w_j, s_j, l_0)\) of the \( j \)th sub-section \((j = 1, 2, \ldots, m)\) are simply given by

\[
\begin{align*}
    w_0 &= W_{\text{line}} \\
    w_j &= W_{\text{line}} + j \cdot \frac{w_{\text{tab}}}{m} \\
    s_j &= S_{\text{line}} - j \cdot \frac{w_{\text{tab}}}{m} \\
    l_0 &= \frac{L_{\text{tab}} - L_{\text{tab,end}}}{2m}
\end{align*}
\]

which will contribute to the S-parameter formulation of each sub-section. According to the known relationship between a \( 4 \times 4 \) S-matrix and ABCD-matrix [106], the ABCD-matrix of each sub-section is converted from corresponding S-matrix by

\[
\begin{bmatrix}
    A_{T1} \\
    B_{T1} \\
    C_{T1} \\
    D_{T1}
\end{bmatrix} = \begin{bmatrix}
    S_{LL} - U_{2 \times 2} & (S_{LL} + U_{2 \times 2})Z_0 \\
    S_{RL}Z_0 & U_{2 \times 2} - S_{RR} & (U_{2 \times 2} + S_{RR})Z_0
\end{bmatrix}^{-1} \begin{bmatrix}
    -S_{LR} \\
    S_{LR}Z_0
\end{bmatrix}
\]

where \( U_{2 \times 2} \) is a \( 2 \times 2 \) identity matrix. As long as \( m \) is large enough, the transitional section T1 could be fully replaced by the stepped structure composed of \( m \) sub-sections cascading together, as observed from Fig. 6.8, leading to

\[
[A_{T1}] = [A_{T1}]_1[A_{T1}]_2 \cdots [A_{T1}]_m
\]

Then, the S-matrix of section T1 \([S_{T1}]\) is converted back from its ABCD matrix \([A_{T1}]\) by
Consequently, Fig. 6.9 compares the S-parameters of the transitional section T1 obtained by the numerical simulation and the method of segmentation in this paper, respectively. If \( m \) is chosen as 1, transition section T1 is actually evolved into a section of asymmetrical coupled microstrip traces with uniform width of \( W_{\text{line}} \), \( W_{\text{tab}} \) and spacing of \( S_{\text{line}} - S_{\text{line}} \).

\[
[S_{T1}] = \left[ \begin{array}{cc}
-U_{2\times2} & A + B/Z_0 \\
U_{2\times2} & CZ_0 + D
\end{array} \right]^{-1} \left[ \begin{array}{cc}
U_{2\times2} & B/Z_0 - A \\
U_{2\times2} & D - CZ_0
\end{array} \right] = \begin{bmatrix}
S11 & S12 & S13 & S14 \\
S21 & S22 & S23 & S24 \\
S31 & S32 & S33 & S34 \\
S41 & S42 & S43 & S44
\end{bmatrix}
\] (6.33)

Figure 6.9 Illustration of comparison for S-parameters of transitional section T1 using numerical simulation and the method of segmentation with \( m = 1, 5, \) and 10, respectively. Its geometrical dimensions are concluded as: \( W_{\text{line}} = 100 \mu m, W_{\text{tab}} = 100 \mu m, S_{\text{line}} = 165.1 \mu m, L_{\text{slope}} = 500 \mu m. \)
It is observed from Fig. 6.9 that the analytically solved S-parameters are gradually approaching the numerical results with the increase of \( m \) (the number of segmentation). There are no obvious differences found when \( m \) is larger than 5, indicating a good accuracy is already achieved. As a conclusion, the S-parameters of equivalent transitional section with segmentation number \( m \) being greater than or equal to 5 could represent those of real transitional section. After that, the S-parameters of section T2, T3, and T4 are separately formulated by:

\[
[S_{T2}] = \begin{bmatrix}
S33 & S34 & S31 & S32 \\
S43 & S44 & S41 & S42 \\
S13 & S14 & S11 & S12 \\
S23 & S24 & S21 & S22
\end{bmatrix} \quad \text{(6.34)}
\]

\[
[S_{T3}] = \begin{bmatrix}
S12 & S11 & S14 & S13 \\
S42 & S41 & S44 & S43 \\
S32 & S31 & S34 & S33 \\
S44 & S43 & S42 & S41
\end{bmatrix} \quad \text{(6.35)}
\]

\[
[S_{T4}] = \begin{bmatrix}
S34 & S33 & S32 & S31 \\
S24 & S23 & S22 & S21 \\
S14 & S13 & S12 & S11
\end{bmatrix} \quad \text{(6.36)}
\]

### 6.4.2 Tab Coupling Fringing Capacitance (TCFC)

![Diagram](image)

**Figure 6.10** Details of the tab-coupling fringing capacitance in the region between adjacent trapezoidal tabs.
It is noticed that in the tab-routing section, there are two types of fringing capacitance that need to be addressed in Fig. 6.10. When the two tabs are interfacing with each other in an interleaving fashion, there are fringing capacitances coming from the top and sides of the trapezoidal shape. There are no literatures that have addressed such fringing capacitances before. Most importantly, omitting such capacitances will definitely degrade the accuracy of the proposed solutions. In this section, we introduce the concept of tab-coupling fringing capacitance (TCFC), namely $C_{TCFC1}$ and $C_{TCFC2}$ in Fig 6.10. $C_{TCFC1}$ is determined by

$$C_{TCFC1} = \frac{C_t - C_{cp}}{2}$$  \hspace{2cm} (6.37)

$$C_t = \frac{1}{c_0 Z_C}$$  \hspace{2cm} (6.38)

$$C_{cp} = \varepsilon_0 \frac{L_{tab\_end}}{S_{line} - W_{tab}}$$  \hspace{2cm} (6.39)

where characteristic impedance $Z_C$ of microstrip line with air as substrate, presented in the inset of Fig. 6.10, is expressed by [106]:

Figure 6.11 Magnified view of the electric field distribution around trapezoidal tabs using commercial tool.
\[
Z_c = \begin{cases} 
60 \ln \left( \frac{8 (S_{\text{line}} - W_{\text{tab}}) \cdot L_{\text{tab,end}}}{L_{\text{tab,end}} \cdot (S_{\text{line}} - W_{\text{tab}})} \right), \\
120 \pi \left[ \frac{L_{\text{tab,end}} / (S_{\text{line}} - W_{\text{tab}}) \leq 1}{\frac{L_{\text{tab,end}} / (S_{\text{line}} - W_{\text{tab}})}{S_{\text{line}} - W_{\text{tab}}} + 1.393} + 0.677 \ln \left( \frac{L_{\text{tab,end}} / (S_{\text{line}} - W_{\text{tab}})}{S_{\text{line}} - W_{\text{tab}}} + 1.444 \right) \right], \\
L_{\text{tab,end}} / (S_{\text{line}} - W_{\text{tab}}) \geq 1
\end{cases} 
\]

(6.40)

Besides, we created an electric field distribution view with the commercial tool in the area of interest, as shown in Fig. 6.11. By observing the electric field distribution in the middle section where two tabs are interleaving, it is noticed that the E-field is not perfectly aligning with the y-direction as expected. As a matter of fact, the tilted electric field is a vector composition of mutual electric field belonging to the middle section and the coupling electric field between the two adjacent tabs.

Therefore, Tab-Coupling Fringing Capacitance (TCFC) \(C_{\text{TCFC2}}\) ought to be considered pertaining to the dimensions and spacing of trapezoidal tabs. As shown in Fig. 6.12, the mutual capacitance \(C_{\text{TCFC2}}\) between tabs in red dashed circle can be approximately obtained by considering that they have trace width of \(L_{\text{tab,end}}\) and spacing of \(L_{\text{mid}}\), and then using previously derived equations Eq. 6.4, 6.9, 6.10 and 6.13, with \(W_1 = W_2 = L_{\text{tab,end}}, s = L_{\text{mid}}\). In the process of generating S-parameters of the middle section, \(C_m\) per unit length belongs to the middle section is modified by adding \(C_{\text{TCFC2}} [W_{\text{tab}} - (S_{\text{line}} - W_{\text{tab}})] / L_{\text{mid}}\) due to extra coupling between adjacent tabs.

With the known S- and ABCD-parameters of each section, the ABCD matrix of the whole structure is theoretically determined by

\[
[A_{\text{Total}}] = [A_1] \cdot [A_1] \cdot [A_M] \cdot [A_2] \cdot [A_M] \cdot [A_N] \cdot [A_0]
\]

(6.41)

where \([A_k]\) \((k = 1, 2, ..., N)\) can be expressed as

\[
[A_k] = [A_{T1}] \cdot [A_{Q1}] \cdot [A_{T2}] \cdot [A_M] \cdot [A_{T3}] \cdot [A_{Q2}] \cdot [A_{T4}]
\]

(6.42)
In the final step, the conversion by adopting Eq. 6.33 is conducted again for deriving the scattering parameter of the complete structure under discussion.

Middle Section

Figure 6.12 Illustration of using approximation method to extract $C_{TCFC2}$ by previously derived equations.

Figure 6.13 Accuracy confirmation of developed analytical equation-based solutions in $S$-parameters for surface tab-routing.
To evaluate the efficacy of proposed analytical equation-based solutions in presenting the S-parameters of the coupled, asymmetrical, lossy and non-uniform microstrip lines with interdigital trapezoidal tabs, we perform numerical simulations based on the structure presented in Fig. 6.6, whose geometrical layout is summarized as: 

$W_{\text{line}} = 400 \, \mu\text{m}$, $W_{\text{tab}} = 304.8 \, \mu\text{m}$, $S_{\text{line}} = 380 \, \mu\text{m}$, $L_{\text{io}} = 330.2 \, \mu\text{m}$, $L_{\text{mid}} = 458 \, \mu\text{m}$, $L_{\text{tab}} = 279.4 \, \mu\text{m}$, $L_{\text{tab\_end}} = 177.8 \, \mu\text{m}$, and $N = 13$. Djordjevic-Sarkar model is applied by defining the frequency-dependent substrate information in ANSYS HFSS. As plotted in Fig. 6.13, the analytical results match the numerical results very well.

6.5 Measurement Results and Discussion

An arbitrary coupled microstrip traces having equal width is provided in Fig. 6.14 (a) for adding interdigital tabs to mitigate far-end crosstalk. As known, it is very time-
consuming to apply ANSYS HFSS for solving the S-parameters of the coupled traces loaded with tabs. Consequently, it is not efficient and practical to use HFSS to perform optimization for tab specifications. However, the proposed closed-form solution is able to find the crucial and optimized parameters that determine the FEXT.

As shown in Fig. 6.14 (b), we start with a configuration with only four tabs incorporated. For instance, the distance between adjacent tabs is controlled to be 2 mm here, which is much larger than the spacing between signal traces, thus leading to negligible mutual coupling between tabs. The influences of parameters $L_{\text{tab}}$ and $W_{\text{tab}}$ on the far-end crosstalk are respectively investigated using the developed equation-based solution. In Fig. 6.15, we can observe and conclude that adding tabs actually contributes little to FEXT reduction when mutual coupling between tabs is negligible. Based on this observation, we further shrink the distance between adjacent tabs by adding more tabs, as shown in Fig. 6.14 (c) and then take the mutual coupling into account by the method given in Part 6.4.2 of section 6.4. After the mutual coupling is introduced, $L_{\text{tab}}$ and $W_{\text{tab}}$

![Figure 6.15](image_url)  

Figure 6.15 Influence of altering (a) $W_{\text{tab}}$ and (b) $L_{\text{tab}}$, as in Fig. 6.14 (b), on far-end crosstalk, respectively.
are fine-tuned to slightly adjust the final S-parameters. Fig. 6.16 compares the S-parameters solved by closed-form equations with their counterparts obtained by commercial tools based on microstrip coupled traces with/without optimized tabs. The good agreement between them fully confirms the merits of our proposed analytical closed-form solutions in optimizing microstrip tab routing. And the optimized dimensions of tabs are concluded as: \( W_{\text{line}} = 400 \mu m, W_{\text{tab}} = 304.8 \mu m, S_{\text{line}} = 400 \mu m, L_{\text{io}} = 560 \mu m, L_{\text{mid}} = 305 \mu m, L_{\text{tab}} = 203.2 \mu m, L_{\text{tab\_end}} = 177.8 \mu m. \)

Figure 6.16  S-parameters comparison of microstrip traces (a) without and (b) with optimized tabs between equation-based solution and commercial tools, respectively.
Figure 6.17  Setup of circuit model to capture the impact of crosstalk on eye diagrams.

Figure 6.18  Comparisons of eye diagrams (a) without and (b) with the optimized tab-routing.
For evaluating the high-speed data transmission performance of the above coupled traces with optimized tabs, a 5-Gb/s PRBS pattern with 50 psec rise/fall time is applied to the transmitter (Tx) on the victim trace, as illustrated in Fig. 6.17. Between the tabbed lines and the terminations, there are single-ended trace sections laid out to mimic the practical routing of DDR buses from CPU to memory chips. The crosstalk (Xtlk) from aggressor trace is set to have the same signal swing at the transmitter but random relative phase $\theta$. The influences of asynchronous (random $\theta$) crosstalk on channel performance are considered and obtained from the eye diagrams. The traditional coupled microstrip lines without tab-routing are also used to generate the eye diagrams for comparison. From Fig 6.18, the surface tab-routing improves eye diagram performance significantly. In the whole application process aforementioned, our equation-based solutions generate the required S4P files in an efficient and effective way without the long computing time needed by the numerical tools. This fast and accurate modeling process can enable an improved eye in the design stages.

6.6 Conclusion

In this chapter, the equation-based solutions are provided for tackling coupled, asymmetrical, lossy and non-uniform microstrip structures. We derive the quasi-static capacitance and inductance matrices for the asymmetrical coupled microstrip lines based on their cross-sectional geometrical dimensions. By converting the capacitance and inductance matrices into frequency dependent causal RLGC parameters, the scattering parameters of uniform coupled lossy microstrip lines with unequal widths are obtained, which are in a good agreement with the corresponding numerical modeling results.
Without loss of generality, the developed approach is also applicable to the case of identical coupled lossy microstrip lines.
CHAPTER 7
DISSERTATION SUMMARY AND FUTURE WORK

7.1 Summary of Contributions

In this dissertation, the contributed work can be divided into two major parts: the design of high performance RF filtering components with multi-band, multi-function, reconfigurable operation for modern wireless communication systems and an accurate equation-based methodology for optimizing the dimensions of closely spaced microstrip tabbed routings which are widely employed in a highly integrated system. Each part has the following contents:

The first part firstly introduce the design of a fourth-order cross coupled single-band filter and a dual-band filter with high frequency selectivity and wide rejection band. The fourth-order cross coupled topology for filter design is investigated by coupling matrix theory. Along with 0° feed structure, multiple finite transmission zeros are created near the passband, thus improving frequency selectivity of designed filters. For extending the rejection band, either adding a lowpass unit or intrinsically staggering the harmonic peaks is adopted and compared. Following that, a multi-functional RF component that fuses the functions of frequency selection and power division is introduced. Two coupled dual-mode resonators incorporating mixed electric and magnetic coupling are used to replace the quarter-wavelength line segments of conventional Wilkinson power divider. The multi-functional component introduced is compact in size compared to that of two
individual components in cascade connection. Afterwards, two RF dual-band filters in balanced architecture are investigated: a balanced dual-band filter with fixed center frequencies, and a balanced dual-band filter with reconfigurable center frequencies. Novel dual-stub loaded resonators are presented as basic building blocks to provide dual-mode resonances at two diverse frequencies under differential-mode excitation. In the end, commercial varactor diodes are embedded into the dual-stub loaded resonators to enable frequency tuning capabilities of developed balanced dual-band filter.

The second part of this dissertation mainly focuses on the accurate equation-based methodology for optimizing microstrip routing traces incorporating interdigital tabs. A set of complete analytical equation-based solutions are developed to predict the behaviors of tabbed routing microstrip lines in a coupled fashion, which has been used to mitigate far-end crosstalk. As an alternative approach to the lengthy numerical simulations, the proposed equation-based solutions are applicable not only to coupled uniform microstrip lines with equal/unequal trace widths but also to the transitional structures with linearly varying trace width. The equation-based solutions are targeted at practical high speed and high-density PCB designs with an optimization process in mind, where the common physical characteristics of transmission lines, such as inhomogeneous, coupled, asymmetrical, lossy, and non-uniform properties, are fully addressed.

7.2 Future Works

Based on the summary of finished work, the future work can be focused on exploring the possibility of making the RF filtering function integrated component reconfigurable. Since a filtering Wilkinson power divider has been developed in Chapter 4 of this dissertation with T-stub loaded dual-mode resonators operating as its basic
building blocks, varactor diodes can be adopted, as shown in Chapter 5 to enable the tunability of developed filtering power divider. In that regard, the proposed reconfigurable filtering Wilkinson power divider will greatly reduce not only the occupied circuit area because the functions of frequency selecting and power dividing are integrated into a single device, but also the manufacturing cost and system complexity because its operation frequency could be reconfigured to different wireless standards.
REFERENCES


