Performance-enhanced microstrip line bandpass filters with size miniaturization

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Performance-Enhanced Microstrip Line Bandpass Filters with Size Miniaturization

Wang Hang

School of Electrical and Electronic Engineering

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in fulfillment of the requirement for the degree of
Doctor of Philosophy

2006

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research done by me and has not been submitted for a higher degree to any other University or Institute.

Nov. -30-2006
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Date

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WANG HANG

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Abstract

The work has been proposed to design various microstrip bandpass filters with miniaturized size, ultra-wide stopband or ultra-wide passband. The stepped impedance resonator (SIR) is used to control the positions of its higher order resonant frequencies by adjusting the impedance ratio. These SIR resonators are then applied to the design of bandpass filters with ultra-wide stopband or ultra-wideband (UWB) passband. The side-coupled microstrip open-loop resonator (OLR) is also studied to remove spurious harmonics by utilizing the coupling-cancellation mechanism. The multi-stage cascaded microstrip bandpass filters constructed on the aperture-backed SIR and side-coupled OLR, dual-mode bandpass filters based on the meander-shaped stepped-impedance ring resonator (SIRR) and UWB bandpass filters on multiple-mode resonator (MMR) have been successfully designed in simulation and verified by experimental measurements. All the newly proposed techniques: aperture-backed SIR, SIRR, MMR, side-coupled OLR and strengthened coupling structures have been found very effective in the enhancement of the filter performance as well as the size miniaturization.

Summary

Planar microstrip bandpass filters with a wide passband have raised great interests in exploring a variety of advanced wireless communication systems. However, the problems of large circuit size and spurious harmonics have been always encountered in filter performances. My PhD work was proposed to solve such problematic issues by presenting various bandpass filters with miniature circuit size, harmonics suppression and wide passband. It is known to all that the microstrip uniform impedance resonator (UIR) always suffers from spurious resonant harmonics at the integral multiples of its first dominant frequency. For this reason, an improved microstrip stepped impedance resonator (SIR) with backed-aperture was developed toward effective suppression of multiple spurious resonances. The low-high-low impedance SIR resonator has a promising capability in achieving an extremely large ratio of the first two resonant frequencies for the design of a bandpass filter with ultra-broad stopband. On the other hand, the high-low-high SIR configuration, referred to as the non-uniform resonator or multiple-mode resonator (MMR), was the essential part in my designed ultra-wideband (UWB) bandpass filter. Based on the theory of orthogonal current distributed characteristic, i.e., opposite even- and odd-symmetrical current distributions in the feed line and the resonator, the side-coupled microstrip open-loop resonator (OLR) provides an alternative good way to remove spurious harmonics.

Firstly, two quarter-wavelength low- and high-impedance transformers were modeled and utilized as alternative types of inductive and capacitive coupling elements with highly tightened degrees for wideband filter design. The three-stage cascaded microstrip bandpass filters were constructed and implemented using the improved SIR and the impedance transformers. Frequency responses over a wide frequency range have quantitatively exhibited their attractive features such as ultra-broad stopband with deep rejection and broadened dominant passband with low insertion loss. Secondly, a novel class of dual-mode bandpass filter was proposed based on the meander-shaped stepped-impedance ring resonator (SIRR). This SIRR consists of low-impedance strips in the four bent corners and high-impedance strips in the four straight sides. A pair of squared notches was then etched out on the corners along either of diagonal lines of this symmetrical SIRR that was capacitively driven by the two perpendicular feed lines. With the cross-shaped aperture placed on the ground underneath the SIRR, the upper stopband was further broadened. Thirdly, harmonic-suppressed bandpass filters with compact size were investigated using the side-coupled microstrip OLR. Two- and four-stage filter blocks were optimally designed with the overall length less than one guided wavelength. Lastly, the UWB bandpass filters have been developed to achieve such a specified UWB passband (3.1-10.6GHz) with a 110% fractional bandwidth at the central frequency of 6.85GHz using the MMRs. The MMRs were constructed with first three resonant modes occurring within UWB band while the aperture-backed parallel microstrip lines or the microstrip/CPW surface-to-surface coupling was largely enhanced. These five-pole UWB bandpass filters were built up and realized with the passband covering the entire UWB

band with low insertion loss, high return loss and small group delay variation. All the newly proposed techniques: aperture-backed SIR, SIRR, MMR, side-coupled OLR and strengthened coupling structures were very effective in the enhancement of the filter performances: broadened stopband or passband as well as in the size miniaturization, which were verified by the commercial software simulations and experimental measurements.

Abbreviations

BPF	bandpass filter
CPW	coplanar waveguide
MSL	microstrip line
MMR	multiple-mode resonator
OLR	open-loop resonator
PCML	parallel-coupled microstrip line
SIR	stepped impedance resonator
SIRR	stepped impedance ring resonator
UIR	uniform impedance resonator
UIRR	uniform impedance ring resonator
UWB	ultra-wideband

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Chapter 1

Introduction

1.1 Background

The word “microwave” is referred to as the electromagnetic waves with wavelength ranging from 1cm to 1m corresponding to the frequency range as high as 30GHz and as low as 300MHz [1]. Generally, the term “microwave engineering” means the engineering and design of information-handling systems in the frequency range from 1GHz up to 100GHz in which lumped circuit elements are usually replaced by transmission line and waveguide components. Several microwave systems such as radar systems, satellites communication, television, radio broadcasting have gained commercial applications.

Microwave filtering devices are the key elements of electronic circuits used in all types of communication systems.

Fig. 1.1 Insertion loss illustrations for ideal case of indoor and handheld systems in [94]

Communication systems require their components such as filters, amplifiers, duplexers and dividers to have high-quality features. With the development of the modern wireless communications and telecommunications, various filters including bandpass, bandstop, highpass and lowpass filters have been always the essential parts in these systems. With respect to the filter design, it requires the filtering device to be low-cost, compact size, light weight and excellent performance within or out of the passband. Little work was done on the design of ultra-wideband (UWB) bandpass filters (BPFs) since the Federal Communications Commission (FCC) released the unlicensed use of the UWB (3.1-10.6GHz) for indoor and hand-held systems (Fig. 1.1) in 2002 [94]. Hence, the existing

issues: harmonics suppression, UWB passband and miniature circuit-size have given a challenge and inspiration to this research work. Due to the limitations of existing structures, novel planar bandpass filters using stepped-impedance resonator (SIR) and open-loop resonator (OLR) need to be developed in order to meet various requirements. These problems are discussed and solved in this thesis after surveying different harmonics suppression techniques, resonators and coupling structures investigated by other people.

1.2 Motivations

As stated in previous section, planar microstrip bandpass filters play an important role in the modern wireless communication and telecommunication systems. However, existing filters require better suppression of unwanted harmonics and reduction of circuit-size. Also, the UWB bandpass filters are becoming one of the challenging topics recently. Therefore the motivations of my PhD work on the filter designs are summarized as follows:

- Higher order resonant modes always occur at the integral multiples of their fundamental resonant frequency in the uniform impedance resonators (UIRs) in addition to their large sizes. These drawbacks inevitably come out when UIRs are cascaded to construct a multistage bandpass filter. Resonators with miniaturized size and improved performance are required.

- Capacitive gap structures usually suffer from weak coupling degree in the design of cascaded multistage bandpass filters with a relative wide passband bandwidth and thus adjustable strengthened coupling structures are demanded.
- Dual-mode bandpass filters using simple ring or patch resonators have attractive features, i.e., small size and sharp out-of-rejection. But little attentions were paid to widening the upper rejection band of the dual-mode bandpass filter.
- Concerning the works on ultra-wideband bandpass filter design, most of them have not met the UWB requirements with satisfaction. In fact, it is difficult to configure such a bandpass filter covering a 110% fractional bandwidth centered at 6.85GHz using the matured microwave filter theory under the assumption of narrow passband.

1.3 Objectives

In accordance with the motivations of the research work, my objectives are designs of size-miniaturized bandpass filters with good harmonic suppression or ultra-wide passband.

The details are listed as follows:

- Firstly, on the design of the stepped impedance resonator, the high-impedance section will be made higher and low-impedance section will be made lower using the aperture compensation technique so that several higher order harmonics are removed.

- Secondly, together with the novel coupling structure having strengthened degree: impedance transformers, the proposed stepped impedance resonators with backed-aperture will be cascaded in the multi-stage bandpass filters to achieve the performance of a wide passband and ultra-wide upper stopband.
- Thirdly, the meander-shaped stepped impedance ring resonator (SIRR) with or without the backed-aperture is to be proposed in order to build up the novel dual-mode bandpass filters with widened upper rejection band.
- Fourthly, multistage bandpass filters using open-loop resonators are to be designed. By properly adjusting the side-coupled feed lines in location, the 2nd resonant frequency is effectively suppressed.
- Lastly, two sorts of UWB bandpass filters will be proposed. The first one is configured on microstrip line structure using aperture-backed parallel-coupled microstrip line and the multiple-mode resonator. The second one is constructed by the hybrid microstrip/CPW coupling and the multiple-mode resonator (MMR) on CPW. All the filters will be designed to cover the UWB passband 3.1-10.6GHz with high return loss and low insertion loss.

1.4 Major Contributions of the Thesis

- A novel improved microstrip stepped impedance resonator: SIR with backed-side aperture, was proposed [119] as a part of this work to greatly increase the spacing between the first dominant resonant frequency and the 1st order harmonic. It has

achieved a size-reduction of 55% compared to its corresponding uniform impedance resonator.

- The microstrip $\lambda/4$ high- and low-impedance transformers have a capacity of tremendously raised coupling extent to be the external coupling of wideband microstrip cascaded bandpass filters. In this way, the proposed aperture-backed SIR and the impedance transformers have constructed a class of bandpass filters [120], [121] with about 10% bandwidth, ultra-broad upper rejection band and miniature circuit size.
- With or without the aperture on the ground plane, the meander-shaped stepped impedance ring resonators applied to the dual-mode bandpass filters [122], [123] have made the 2nd resonant frequency be three to four times of the dominant frequency in addition to the great size miniaturization.
- The 2nd resonant passband suppression technique in the bandpass filter using bent-shape open-loop resonator was originally proposed [124] and it was realized by adjusting the overlapped length of the side-coupled feed lines.
- Two classes of UWB bandpass filters were proposed and implemented [125]-[128] to achieve the five-transmission-pole UWB passband from 3.1GHz to 10.6GHz with high return loss and low insertion loss. As only a single resonator was utilized, the overall lengths of the filters were all about one-wavelength at the central UWB frequency.

Entirely speaking, several novel bandpass filters have been presented to successfully suppress spurious harmonics or realize the UWB passband. They are all small in size. These objectives are achieved by the techniques of SIR with backed-aperture, impedance transformers, meander-shaped SIRR, side-coupled OLR, parallel coupled microstrip lines and hybrid microstrip/CPW coupling structure.

1.5 Organization of the Thesis

In addition to the introduction presented in the first chapter, the following contents from Chapter 3 to Chapter 7 are the own work done and they are organized as follows:

Chapter 3 mainly deals with the characteristic analysis of the improved microstrip stepped impedance resonator after the overview of microwave bandpass filters given in Chapter 2. Three types of $\lambda/2$ SIRs with low-impedance in two open ends and high-impedance in the center have a potential in widening the spacing between the dominant frequency and the 1st spurious resonance as long as the impedance ratio is properly adjusted.

Design, simulation and fabrication of two novel three-stage bandpass filters are described in Chapter 4 using the optimized aperture-backed stepped impedance resonator and the impedance transformers. The proposed high or low impedance transformer, characterized as J- or K-inverter in equivalent filter blocks, greatly strengthen the coupling degree between the feed lines and the first/last stage resonators. As an evident verification, the

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fabricated circuits are measured to showcase that they have a relative wide bandwidth, ultra-broad upper stopband and size reduction.

Chapter 5 investigates the dual-mode bandpass filter using the stepped impedance ring resonator. The perpendicularly oriented feed lines and perturbation elements are essential in the dual-mode bandpass filter construction. As the slit perturbation is found not good at harmonic suppression, the square notch perturbation on corners of the novel meander-shaped SIRR is presented. This novel structure not only makes the two orthogonal modes sufficiently coupled but also remains the harmonic suppression characteristic of the original SIRR.

Chapter 6 presents the bandpass filters with open-loop resonator side-coupled to the bent feed lines. Similarly to the resonant modes existing in the previously discussed ring resonator or closed-loop resonator, the even- and odd-symmetrical current distributions in the feed lines and the first/last stage open-loop resonator take effect on suppressing the 2nd resonant frequency. Two-stage and four-stage bandpass filters are constructed and designed, exhibiting a fairly good passband with high return loss and low insertion loss. Moreover, the 1st harmonic proves to be greatly suppressed as long as the feed line length is properly chosen.

Chapter 7 focuses on the design of the ultra-wideband bandpass filter. In the first class of UWB bandpass filter, the parallel-coupled microstrip line with backed-aperture is used

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together with the microstrip multiple-mode resonator. In the second class of UWB bandpass filter, the hybrid microstrip/CPW coupling structure and CPW non-uniform resonator are utilized. All of the UWB bandpass filters have achieved the FCC-specified UWB passband with low insertion loss, high return loss and flat group delay.

As the last chapter, Chapter 8 concludes the whole work. At last, it should be mentioned that all the design bandpass filters in my work were simulated by ADS Momentum software and measured by Agilent network analyzer.

Chapter 2

Microwave Bandpass Filters

2.1 Introduction

A microwave filter is able to transmit some desired signals in a certain passband or attenuate undesired ones in some other bandwidth. In general, microwave filters are categorized into four types: low pass, high pass, bandpass and bandstop filters. Fig. 2.1 (a)-(d) describes ideal frequency responses of these four types in terms of the transmission coefficient $T(\omega)$. The lowpass filter transmits all the signals below the cutoff frequency ω_c in Fig. 2.1(a) and rejects all above ω_c while the highpass filter in Fig. 2.1(b) rejects all the signals below ω_c and passes all above ω_c . As illustrated in Fig. 2.1(c) and (d), the

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bandpass and bandstop filters only make the signals between ω_{c1} and ω_{c2} transmitted or rejected.

Fig. 2.1 Descriptions of four types of filter
(a) Low pass filter (b) Highpass filter (c) Bandpass filter (d) Bandstop filter

To achieve harmonics suppression, the photonic bandgap (PBG) or other periodic structures [2]-[4] and the defected ground structure (DGS) [5]-[9] have been introduced in the microstrip low pass filters or antennas. Photonic bandgap structure was firstly studied in optical region and has been applied to the microwave field in order to construct wide stopband. The defected ground structure was realized by cutting out the metal on the ground and it was used more than the PBG in filters, amplifiers and dividers. Novel implementations like periodic grooves [10], wiggly-line [11]-[14] in similar periodic form

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using continuous perturbations of coupled-line were also investigated. The wiggly-coupled line has the ability to equate the phase velocities of the odd-mode and even-mode so that the second spurious harmonic is canceled out. In [19], [20], the tapped excitation technique was developed to improve the out-of-passband characteristics, thus suppressing the higher-order spurious harmonics. The over-coupled filter together with the utilization of even- and odd-mode phase constants [15] have been proposed, exhibiting an effective way in significantly suppressing the unwanted responses to lower than -50dB . Also, many other methods were used such as the applications of transmission poles and open stubs [16]-[18], semi-loop shunt arm [21], split ring resonators [22], tapped resonator with dielectric rods [23], high-impedance transmission line with capacitors [24] and the combination of the low pass filters and high pass filters configured on coplanar waveguide [25] to enhance the in-band and out-of-passband characteristics.

There are different kinds of resonator elements used in multi-stage filters cascaded in line [26]-[28] or cross-coupled to each other [29]-[32]. The so-called hairpin resonator filters occupy relative small circuit areas as the resonators were bent in shape and cross-coupled to each other. This novel cross-coupled configuration was widely proposed in the filter designs of Chebyshev, elliptic and quasi-elliptic function [33]. Of many methods that have been mentioned and referenced in harmonics suppression, the stepped impedance resonator (SIR), as originated in stripline filters [34] and further analyzed in [35]-[37], has gained a wide application in exploring a variety of advanced circuit systems like power divider [38], diplexers[39] and bandpass filters based on coplanar waveguide [28] and

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microstrip configurations [40]-[43]. In [34], the half-wavelength stepped impedance resonator was firstly proposed for the design of the parallel coupled bandpass filters. In addition, the authors in [35]-[37] introduced quarter- and one-wavelength type in a complete overview and discussed about a series of this microstrip SIRs and their applications. Therefore, various microstrip cascaded multistage or dual-mode BPFs based on the stepped impedance resonator configuration are supposed to have the advantages of spurious harmonic suppression and compact size.

As a coupling structure in the cascaded filters, the strip-to-strip coupling or the capacitive gap [44] structure was usually placed between cascaded stages. Together with some other mechanisms such as multi-layer electric coupling aperture [45] and interdigital capacitor [46], it was used as the application of relative narrow-band design. In order to devise a relative wider pass bandwidth, the gap is not an adequate coupling method as the external coupling between the feeding line and the first/last stage of resonator. In the CPW bandpass filter, the asymmetric parallel-coupled stage [47]-[49] and high-impedance transformer [50], [51] have provided frequency-dispersive and enhanced couplings to a great extent, showing that they were good techniques in the wider passband design.

Besides, the symmetrical patch or ring resonators are very popular in dual-mode bandpass filters due to their low radiation loss feature and capability in elimination of the parasitic resonances existing in the open- or short-ended quarter- and half-wavelength resonators. In the past several years, various microstrip dual-mode bandpass filters have been

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extensively studied and implemented [37], [52]-[81] in the form of circular patch or ring [37], [53]-[55], [59], [60], [70], [73]-[76], square patch or ring [56], [58], [61]-[69], [71], [77]-[79] and triangular patch or ring [57], [80]. The concept of a microstrip ring dual-mode filter was originally proposed in [52] by setting up the two perpendicular feed lines and exciting the two degenerate resonant modes on an annular uniform impedance ring resonator (UIRR). By attaching the tuning stub perturbations on the diagonal line that is symmetrical with respect to the two feed lines, the UIRR-based dual-mode ring filters were developed and designed using the simple transmission line theorem [53]-[57]. Recently, a capacitively-loaded slow-wave ring resonator filter with squared shape [58] and open-loop meandered line perturbation [62], [63] were reported to reduce the overall size and broaden the upper stopband. The popularly used defected ground structure [59]-[61] and EBG structure [64] also have the size-miniaturization and harmonic suppression function in the design of the dual-mode filters. In the meantime, the work in [55], [65]-[67] discussed how the transmission zeros could be produced and appropriately allocated on the lower and upper rejection skirts outside the passband by constructing the inductive or capacitive perturbation in a proper diagonal line of a ring or patch resonator. The rejection skirts with or without transmission zeros were also investigated in the cross-slotted patch dual-mode filters with inductive [68] or capacitive loading [69]. In order to enhance the external coupling of the dual-mode bandpass filter, the side-line coupling [58], [62], [63], [69] between the feeding line and resonator in terms of side or direct coupling [70], [71] was commonly used. In [72], a meander-shaped UIRR was presented to build a

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size-reduced dual-mode ring filter and this meander line structure gave the space to tighten the line-to-ring coupling [53], [69].

Microstrip open-loop resonators (OLRs) were used in developing a new type of cross-coupled planar bandpass filters with elliptic-function [83] or quasi-elliptic function [84] response. Further investigations on the square open-loop resonator filter appeared in [85] with a single pair of attenuation poles at finite frequencies based on design tables and formulas for accurate and efficient filter synthesis. Originated in [86], a multistage cascaded open-loop resonator filter was proposed by tapping the input and output feed lines to the asymmetrical position of the first and last open-loop resonators [87]. In this way, the two transmission zeros can be simultaneously excited and properly allocated on the lower and upper sides of the dominant passband, thus leading to the sharp cutoff frequency response. Using the concept of a stepped impedance resonator in [34], a novel slow-wave open-loop resonator with capacitive loading at two ends was invented [88] to make a new class of cross-coupled bandpass filters with compact size and wide stopband.

Moreover, significant research activities and interests have been seen [95], [96] in UWB application recently. For example, UWB antennas [97] and amplifiers [98], [99] have been explored by many researchers. Actually a few years ago, some wideband bandpass filters with bandwidth not exceeding 70% were proposed [100]-[102]. In [100], [101], the parallel-coupled line bandpass filter was constructed using tightly-coupling three line structure but the reported bandwidth only achieved 40-70%. By means of two stopbands

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of a filter block with the help of two tuning stubs on a ring resonator, a bandwidth of 49.3% was obtained in [102]. As one of the essential component blocks, an UWB bandpass filter (BPF) has tried to be developed in [103]-[109] to achieve such a specified UWB passband with a 110% fractional bandwidth at the central frequency of 6.85GHz. In [103], an initial UWB filter was presented by mounting a microstrip line in a lossy composite substrate and the reported insertion loss is higher than 6.0dB. In [104]-[105], a microstrip ring UWB filter was constructed by simultaneously exciting and allocating transmission zeros below 3.1GHz and above 10.6GHz. Due to its nature of dual-stopband, this filter with multiple ring resonators usually has narrow lower and upper stopbands as well as large size. In [106], a called composite UWB filter was proposed by combining lowpass and highpass filter structures or embedding one into the other. In [107], a broadside-coupled microstrip-coplanar waveguide (CPW) structure with tightened coupling degree was utilized to design an alternative UWB filter with one, two and three sections. An extremely wideband filter using ridge waveguide [109] was presented with a very wide bandwidth near the UWB requirements, but had a structure disadvantage of the complex transition and 3D layout. In [108], a novel compact UWB bandpass filter on microstrip line was constituted using a single multiple-mode resonator (MMR) that was driven at two sides by two identical parallel-coupled lines. The basic principle of this UWB filter was originated in [110]-[112] to explore compact and broadband bandpass filters with the bandwidth of 60%~ 80%. Other than the parallel-coupled microstrip lines in [110]-[112], another coupling mechanism called surface-to-surface coupling [113]-[115] by means of microstrip-to-CPW [113], [115] or microstrip-to-microstrip transition [114]

has realized a flat frequency response of low insertion loss over a very wide bandwidth up to 111% in [115], indicating that this back-to-back coupling technique was a good candidate for UWB filter design as well.

It is seen that SIR, SIRR, MMR and OLR have their advantages in the performance-enhanced bandpass filters with ultra-wide stopband or passband. As stated in Table 2.1, they have been realized with small size and will be detailed in my work.

Table 2.1 Performances of the proposed resonators in filter design

	Size Reduction	Harmonic Suppression	UWB Band
SIR in Cascaded BPF	Yes	Yes	/
SIRR in Dual-mode BPF	Yes	Yes	/
MMR in UWB BPF	Yes	/	Yes
OLR in Cascaded BPF	Yes	Yes	/

2.2 Bandpass Filter Topologies

The microwave bandpass filter can be presented in terms of equivalent LC network with K- or J-inverter. Fig. 2.2 shows the ladder network of a microwave bandpass filter. This initial filter requires both series and parallel LC resonators, but it seems difficult in the simultaneous implementation of these two types of LC resonators. Therefore, the immittance inverter, i.e., impedance or admittance inverter is usually utilized to build up a filter block with either series or parallel elements.

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Fig. 2.2 Ladder network of the microwave bandpass filter

Fig. 2.3 Definition and block diagram (a) Impedance inverter (b) Admittance inverter

Fig. 2.3 (a) and (b) show the two-port impedance and admittance inverters that invert the load impedance Z_b or admittance Y_b to the input Z_a or Y_a in terms of the characteristic impedance or admittance of the square block named K or J respectively in (2.1):

Fig. 2.4 Shunt admittance and series impedance transformation (a) K-inverters (b) J-inverters

$$Z_a = \frac{K^2}{Z_b}, Y_a = \frac{J^2}{Y_b} \quad (2.1)$$

Considering the shunt admittance $Y_p(\omega)$ with two side K-inverters in Fig. 2.4(a), the two-port network can be taken as $Y_p(\omega)$ in parallel: $Z_s(\omega)$ in series with the two open circuits at the two sides if the output ports are both terminated with short circuits. With respect to the series impedance $Z_s(\omega)$ with J-inverters in Fig. 2.4(b), it is equivalent to the shunt admittance $Y_p(\omega)$ [1].

Fig. 2.5 Equivalent networks of lumped element microwave bandpass filter
(a) Impedance inverter (b) Admittance inverter

The unique feature of these immittance inverters enables us to convert the practical filter block to those with either series or parallel elements. According to the above analysis, the ladder network in Fig. 2.2 can be modified to Fig. 2.5(a) including only the series elements or Fig. 2.5(b) including only the shunt elements. These filter blocks give us a convenient way to devise the microwave bandpass filter with only series or parallel

lumped/distributed resonators. In the filter design of the thesis, the LC resonators are realized by the stepped-impedance resonator or open-loop resonator, whose principle is discussed in the following sections. And the K-/J-inverters by means of different coupling structures are detailed in later respective chapters.

2.3 Microstrip Stepped Impedance Resonator

As described by the authors in [34] and [35], Fig. 2.6 (a), (b) and (c) depict the layouts of three typical kinds of stepped impedance resonators: quarter-, half- and one-wavelength SIR, in which all of the SIRs have two distinctive sections with characteristic impedance Z_1 , Z_2 and respective electrical length θ_1 and θ_2 .

Fig. 2.6 Structure of the typical SIR
(a) Quarter-wavelength type (b) Half-wavelength type (c) One-wavelength type

2.3.1 $\lambda/4$ and $\lambda/2$ SIR

Fig. 2.7 illustrates the $\lambda/4$ SIR structure with open and short ends at the two sides and its equivalent transmission line circuit. The input impedance from the left side in Fig. 2.7(b) is defined as Z_i and $Y_i=1/Z_i$ is the input admittance. The equivalent circuit of this resonator can be considered as two parts of transmission line with different characteristic impedance Z_1 and Z_2 . According to the equation (2.2) in [1], the input admittance Y_i of the cascaded transmission line with the short circuit load as $Z_L=0$ can be derived as indicated in equation (2.3) [35].

Fig. 2.7 Resonant condition of $\lambda/4$ SIR
(a) $\lambda/4$ SIR layout (b) Equivalent transmission line topology

$$Z_{in} = Z_0 \frac{Z_L + j Z_0 \tan \theta}{Z_0 + j Z_L \tan \theta} \quad (2.2)$$

$$Z_i = j Z_2 \frac{Z_1 \tan \theta_1 + Z_2 \tan \theta_2}{Z_2 - Z_1 \tan \theta_1 \tan \theta_2} = j Z_2 \frac{\tan \theta_1 + K \tan \theta_2}{K - \tan \theta_1 \tan \theta_2} \quad (2.3)$$

where K is the impedance ratio $K=Z_2/Z_1$. The layout of the half-wavelength SIR and its equivalent circuit are illustrated in Fig. 2.8. The input admittance Y_i is treated in a similar

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way of the $\lambda/4$ SIR except that it has three cascaded sections and the open-circuited load of $Y_L=0$. So Y_i is expressed as follows [35]:

$$Y_i = jY_2 \frac{2(K \tan \theta_1 + \tan \theta_2)(K - \tan \theta_1 \tan \theta_2)}{K(1 - \tan^2 \theta_1)(1 - \tan^2 \theta_2) - 2(1 + K^2) \tan \theta_1 \tan \theta_2} \quad (2.4)$$

Fig. 2.8 Resonant condition of $\lambda/2$ SIR
(a) $\lambda/2$ SIR layout (b) Equivalent transmission line topology

The resonant condition must satisfy that $Y_i=1/Z_i=0$ at either open end such that [35]:

$$Z_2 = Z_1 \tan \theta_1 \tan \theta_2 \quad (2.5)$$

$$K = Z_2 / Z_1 = \tan \theta_1 \tan \theta_2 \quad (2.6)$$

In the other possibility of $K = -\tan \theta_2 / \tan \theta_1$, either $\tan \theta_2$ or $\tan \theta_1$ is required to be smaller than zero for the reason of $K > 0$ and then we have $\pi/2 < \theta_1$ or $\theta_2 < \pi$. The total phase or electrical length ($2\theta_1 + 2\theta_2$) of this SIR becomes much larger than π , thus implying the 2nd- or high-order resonant mode case. Therefore the possibility of $K = -\tan \theta_2 / \tan \theta_1$ is not taken into account. Unlike the uniform impedance resonator resonance condition that is mainly related to its overall electrical length, the above equations give us an idea that the resonance condition of SIR is not determined by one single aspect but more than one: the impedance ratio K and the respective electrical length of the two segments. The total electrical length of the quarter-wavelength SIR, θ_A , can be expressed as follows:

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$$\theta_A = \theta_1 + \theta_2 = \theta_1 + \arctan\left(\frac{K}{\tan \theta_1}\right) \quad (2.7)$$

So the total electrical length of the half-wavelength and the one-wavelength SIR is two and four times of θ_A , i.e. $2\theta_A$ and $4\theta_A$. When it comes to the normalized resonator length L_n , SIR is associated with the UIR electrical length $\pi/2$, π and 2π and therefore:

$$L_n = \frac{\theta_A}{\pi/2} = \frac{2\theta_A}{\pi} = \frac{4\theta_A}{2\pi} \quad (2.8)$$

It can be seen from equation (2.8) that the normalized resonant lengths are equal to each other in the cases of $\lambda/4$ -, $\lambda/2$ - and λ -type resonators. According to equation (2.6) and (2.7), substituting $\theta_2 = \theta_A - \theta_1$ to (2.6):

$$K = \tan \theta_1 \tan(\theta_A - \theta_1) = \tan \theta_1 \frac{\tan \theta_A - \tan \theta_1}{1 + \tan \theta_A \tan \theta_1} \quad (2.9)$$

$$\text{So, } \tan \theta_A = \frac{K + \tan^2 \theta_1}{(1 - K) \tan \theta_1} = \frac{1}{1 - K} \left(\frac{K}{\tan \theta_1} + \tan \theta_1 \right) \quad (2.10)$$

When $0 < K < 1$ ($1 - K > 0$), we can obtain $0 < \theta_A < \pi/2$

$$(\theta_A)_{\min} = \arctan\left(\frac{2\sqrt{K}}{1 - K}\right) \quad (2.11)$$

When $K > 1$ ($1 - K < 0$), $\pi/2 < \theta_A < \pi$ is valid

$$(\theta_A)_{\max} = \arctan\left(\frac{2\sqrt{K}}{1 - K}\right) \quad (2.12)$$

θ_A reaches the maximum or minimum value when the condition meets that:

$$\theta_1 = \theta_2 = \arctan(\sqrt{K}) \quad (2.13)$$

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(a)

(b)

Fig. 2.9 Normalized resonant length versus K (a) $0 < K < 1$ case (b) $K > 1$ case

Fig. 2.9 illustrates the normalized resonator length versus the impedance ratio K . In the first case of $0 < K < 1$ in Fig. 2.9 (a), other curves are all above solid line under the condition

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when $\theta_2 = \theta_1$. The curve under $\theta_1 = \pi/6$ is very close to the solid line at higher K and the curve under $\theta_1 = \pi/12$ is close to it at lower range of K . In the second case of $K > 1$ in Fig. 2.9(b), all the curves are below that of the condition $\theta_2 = \theta_1$ except that $\theta_1 = 5\pi/12$ dotted-line close to it at larger K and $\theta_1 = \pi/3$ dashed-line close to it at smaller K . These results intuitively tell us that a smaller normalized resonator length arises at some specific impedance ratio as long as the two electrical lengths: θ_2, θ_1 are properly chosen.

Referring to the analysis in [35], the first spurious harmonic in $\lambda/4, \lambda/2, \lambda$ -type SIR are set to be f_{AI}, f_{BI} and f_{CI} respectively and the normal case of $\theta_1 = \theta_2 = \theta_0$ is assumed in the following discussions. For the $\lambda/4$ case, (2.6) leads to,

$$\tan^2 \theta_0 = \tan^2 \theta_{SA1} = K \quad (2.14)$$

$$\text{So, } \theta_0 = \arctan(\sqrt{K}) \quad (2.15)$$

$$\theta_{SA1} = \arctan(-\sqrt{K}) + \pi = \pi - \theta_0 \quad (2.16)$$

The θ_{SA1} is the electrical length corresponding to the first spurious resonant frequency and the others should be $\theta_{SA n} = \pi + \theta_0, 2\pi - \theta_0, 2\pi + \theta_0 \dots$ ($n=2,3,4 \dots$)

$$\frac{f_{SA1}}{f_0} = \frac{\pi - \theta_0}{\theta_0} = \pi / \arctan(\sqrt{K}) - 1 \quad (2.17)$$

$$\frac{f_{SB1}}{f_0} = \frac{f_{SC1}}{f_0} = \frac{\pi + \theta_0}{\theta_0} = \pi / \arctan(\sqrt{K}) + 1 \quad (2.18)$$

For the $\lambda/2$ -, λ -type cases (2.4) becomes,

$$Y_i = jY_2 \frac{2 \tan \theta_0 (1 + K)(K - \tan^2 \theta_0)}{K - 2(1 + K + K^2) \tan^2 \theta_0 + K \tan^4 \theta_0} = 0 \quad (2.19)$$

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As illustrated in formula (2.19), $\lambda/2$ -, λ -type SIR have the same θ_0 with $\lambda/4$ SIR and all the other roots of (2.19) are the electrical length satisfying the multiple resonant frequencies and they can be determined as,

$$\theta_{SB1} = \theta_{SC1} = \pi / 2 \quad (2.20)$$

$$\theta_{SB2} = \theta_{SC2} = \pi + \arctan(-\sqrt{K}) = \pi - \theta_0 \quad (2.21)$$

$$\theta_{SB3} = \theta_{SC3} = \pi \quad (2.22)$$

Fig. 2.10 Relationship among the f_0 and higher order harmonics f_1, f_2, f_3

Therefore:

$$\frac{f_{SB1}}{f_0} = \frac{f_{SC1}}{f_0} = \pi / 2 \arctan(\sqrt{K}) \quad (2.23)$$

$$\frac{f_{SB2}}{f_0} = \frac{f_{SC2}}{f_0} = \frac{\pi - \theta_0}{\theta_0} = \pi / \arctan(\sqrt{K}) - 1 \dots \quad (2.24)$$

In Fig. 2.10, the dominant frequency f_0 of 1GHz, as well as the higher order harmonics f_1 , f_2 , and f_3 , are depicted together versus the impedance ratio K . As K decreases from 1 to 0, the spurious resonant frequencies can be greatly shifted to the higher range against f_0 . When K is larger than 1, the curves are flat in shape and f_1 , f_2 and f_3 are approaching to f_0 . The first type SIR of $0 < K < 1$ is mainly used in the design of harmonic suppression filters with the improved upper rejection band. On the other hand, the higher order resonant modes, closely spaced in the second type SIR of $K > 1$, may be utilized to construct very wideband multiple-mode bandpass filters like UWB bandpass filters.

2.3.2 Stepped Impedance Ring Resonator (SIRR)

Fig. 2.11 (a) and (b) depict the layouts of the traditional squared uniform and stepped impedance ring resonators, i.e., UIRR and SIRR, which are driven in parallel by the two feed lines via weakly capacitive coupling gaps. In the proposed SIRR, the two feeding lines are capacitively coupled to the narrower microstrip line with wider strip conductors at four corners. It should be pointed out that this SIRR is different from the layout in Fig. 2.6 (c) and designed in this way to be applied to the dual-mode bandpass filters of my work. According to the theorem reported in [34], this SIRR can be expected to effectively control the stopband between the fundamental and 1st spurious resonant frequencies, i.e., f_0 and f_1 , in terms of the impedance ratio of the narrow and wide strip sections in Fig. 2.11 (b).

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Fig. 2.11 Layouts of two types of ring resonator (a) UIRR (b) SIRR A

Fig. 2.12 Static capacitance-based network topology for the stepped impedance ring resonator

Fig. 2.12 presents the equivalent network of the stepped impedance ring resonator in Fig. 2.11 (b), in which the capacitive gap coupling is characterized by a simple capacitance. To determine all the resonant frequencies, one needs to derive the input admittance, Y_i , at the input or output feeding point of this ring resonator. Considering the geometrical symmetry of the ring resonator with respect to the plane, P-P', this input admittance, Y_i , can be expressed in the even and odd symmetrical cases. For the even-symmetrical case, the P-P' plane is the open-circuit such that:

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$$Y_{i,e} = jY_2 \frac{2(K \tan \theta_1 + \tan \theta_2)(K - \tan \theta_1 \tan \theta_2)}{K(1 - \tan^2 \theta_1)(1 - \tan^2 \theta_2) - 2(K^2 + 1) \tan \theta_1 \tan \theta_2} \quad (\text{For even-mode}) \quad (2.25)$$

For the odd-symmetrical case, the P-P' plane represents the short-circuit such that:

$$Y_{i,o} = jY_2 \frac{2(K^2 + 1) \tan \theta_1 \tan \theta_2 - K(1 - \tan^2 \theta_1)(1 - \tan^2 \theta_2)}{2(\tan \theta_1 + K \tan \theta_2)(1 - K \tan \theta_1 \tan \theta_2)} \quad (\text{For odd-mode}) \quad (2.26)$$

where K is the impedance ratio Z_2/Z_1 . Equation (2.25) and (2.26) can be used to solve the resonant frequencies of even- and odd-modes, i.e., $f_1, f_3, f_5, \dots$ and $f_0, f_2, f_4, \dots$, respectively.

The input admittance must be equal to zero in order to meet the resonance condition. In the case of odd-mode, we have:

$$(1 - \tan^2 \theta_1)(1 - \tan^2 \theta_2) - 2 \tan \theta_1 \tan \theta_2 (K + 1/K) = 0 \quad (2.27)$$

So,

$$\tan 2\theta_1 \tan 2\theta_2 = \frac{2K}{K^2 + 1} \quad (2.28)$$

Assuming that $\theta_1 = \theta_2 = \theta_0$ in odd-mode, one can further obtain,

$$\theta_o = \frac{1}{2} \arctan(\sqrt{K'}) \pm \frac{n\pi}{2} \pm \frac{1}{2} \arctan(\sqrt{K'}) \quad (K' = \frac{2K}{K^2 + 1}) \quad (n=1,2,3,\dots) \quad (2.29)$$

According to Equation (2.25), the electrical length for even-mode frequency becomes under the same assumption ($\theta_1 = \theta_2 = \theta_e$):

$$\theta_e = \arctan \sqrt{K}, n\frac{\pi}{2}, n\pi \pm \arctan \sqrt{K} \quad (n=1,2,3,\dots) \quad (2.30)$$

Based on Equations (2.29) and (2.30), the first four resonant frequencies f_0, f_1, f_2 and f_3 can be determined. Fig. 2.13 describes the ratios of the first three higher order resonant frequencies f_1, f_2 and f_3 over the first dominant resonant frequency f_0 with respect to the

impedance ratio K . It is found that the higher order resonant frequencies are shifted away from f_0 when K is largely increased from 1.0 to 8.0. The results also validate that the higher order resonant frequencies occur at the integral multiples of the 1st one in the uniform impedance ring resonator ($K=1$) as the dash-dotted lines are shown in Fig. 2.13.

Fig. 2.13 Resonant frequency ratio f_n/f_0 versus impedance ratio $K=Z_2/Z_1$

2.4 Side-Coupled Open-Loop Resonator

Fig. 2.14 depicts the geometry of the microstrip open-loop resonator side-coupled to an open-end feed line. The overall open-loop resonator has the length of $2L$. The 1st resonance of this open-loop happens at the frequency (f_0), where $2L=\lambda_{g0}/2$ [82]. Intuitively, its 2nd resonance approximately appears at $2f_0$ ignoring the effects of bend discontinuity and frequency dispersion. This spurious resonance unfortunately causes the 1st harmonic passband in the multistage filter design [83, 87]. Based on the discussion on the mode chart in [82], it can be understood that the standing-wave current density along the open-loop at $2f_0$ is odd-symmetrically distributed. It implies that the central location of

this loop is actually a perfect open circuit, marked by “x”, in this case.

Fig. 2.14 Geometry of the considered side-coupled microstrip open-loop resonator with distinct current distributions at $2f_0$

As illustrated in Fig. 2.14, the open-end microstrip feed line is put in close proximity to the middle section of this open-loop. Its relevant line length (D) is referred to as the route from the open end to the center of the parallel-coupled section as shown in Fig. 2.14. The coupling between the feed line and open loop can be effectively controlled by selecting this length (D) according to the distributed coupled line theorem, e.g., [15]. In order to cancel this 2nd resonance at $2f_0$ of the open-loop resonator, the length (D) needs to be properly selected as $L/2$, i.e., $D = L/2 \approx \lambda_{g0}/8$. As such, the even-symmetrical current distribution can be established along the feed line with the perfect short circuit at the vertical center as shown in Fig. 2.14. This orthogonally distributed characteristic enforces no energy transfer between the feed line and open-loop resonator, thus canceling this 1st spurious resonance at $2f_0$.

2.5 Conclusion

The performances of bandpass filters need to be improved in the upper stopband as well as ultra-wideband bandpass filters are demanded in various applications. In this chapter, the characteristics of stepped impedance resonators and open-loop resonators which are investigated in this work have been briefly discussed. Unlike the uniform impedance resonator that has spurious harmonics occurring at the multiple integrals of the fundamental frequency (f_0), the SIR resonator can make the higher resonant frequencies close to or away from f_0 . This SIR configuration is suitable in the design of cascaded or dual-mode bandpass filters with ultra-stopband or UWB passband by adjusting the impedance ratio and the physical/electrical length of the respective parts. Also, the open-loop resonator with adjustable length of open-end feed line is effective to achieve the harmonic suppression in filters.

Chapter 3

Aperture-Backed SIR

3.1 Introduction

As mentioned previously, microstrip line resonators were considered and used as one of the most basic elements in the design of various bandpass filters. Among so many methods in overcoming spurious resonances, the stepped impedance resonator, as originated in stripline filters [34]-[36], has gained a wide application in exploring a variety of advanced filter circuits. In [40], a SIR microstrip filter with tapped-line excitation at the input and output was developed to further extend the upper rejection band. By properly allocating the two transmission zeros, the first and/or second spurious resonances in this filter were effectively cancelled. In [26], by forming an improved SIR

with folded high-impedance section at center, the $2f_0$ harmonic resonance was completely eliminated and the first spurious resonance was raised close to $3f_0$. In this chapter, a few structures of the microstrip SIR are presented for the further design of the novel aperture-backed SIR [119]. After simulations on various SIR configurations, the 1st spurious resonance of the improved SIR with backed aperture [116] underneath central strip section, which is constructed on RT/Duroid 6010 substrate with the permittivity of $\epsilon_r=10.8$ and thickness of $h=0.635\text{mm}$, has been pushed up to higher than $4f_0$ in both simulation and experiment.

3.2 High-Impedance Microstrip Line

3.2.1 Microstrip Line Characteristics

The microstrip is easy to be implemented as a typical transmission line model. Fig. 3.1 (a) illustrates the three-dimensional structure of a microstrip line and Fig. 3.1 (b) is its cross section view. It is composed of a strip conductor with width W placed on the dielectric substrate with a permittivity ϵ_r and height of H . Because microstrip line is not totally surrounded by the dielectric substrate and its upper area is filled with air or free space, the fundamental propagation mode is not a pure TEM mode but so-called quasi-TEM mode at low frequencies of about a few gigahertz. In the frequency range of a few gigahertz or a little lower, the microstrip transmission line can be characterized in terms of the distributed parameters: per-unit-length capacitance and inductance. Fig. 3.1 (b) shows the electric field existing in the microstrip line structure. At higher frequencies, the electric

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field is more confined to the region between the microstrip and ground plane compared with the case at lower frequencies. When the aperture on the ground is introduced in Fig. 3.1(c), more electromagnetic energy is coming out of the region between strip and ground plane. This causes the changes of the effective dielectric constant ϵ_{re} as well thus affecting the electrical length of the transmission line.

Fig. 3.1 Microstrip line (a) 3-D Layout (b) Electric field distribution (no aperture) in cross-section (c) Electric field distribution (with aperture) in cross section

The backed aperture or reduced strip width disperse the distribution of electromagnetic field thus resulting in the coupling decrement between the strip and ground plane, i.e., increasing transmission line distributed capacitance L and decreasing inductance C . As the lossless transmission line, the characteristic impedance Z_0 is expressed in the form of L and C :

$$Z_0 = \sqrt{\frac{L}{C}} \quad (3.1)$$

Therefore, the characteristic impedance Z_0 is greatly increased by further narrowing the upper strip width and etching out aperture on the bottom due to the reduction of the strip-to-ground coupling and the enhancement of the transmission line inductance. Likewise, the wider strip results in the higher distributed capacitance and lower distributed inductance, thus deriving a small Z_0 .

3.2.2 De-embedding of Microstrip Line with Backed-Aperture

The microstrip line with the backed-aperture on the ground has been characterized and its two-port S-parameters have been simulated by ADS Momentum software in the course of

Fig. 3.2 Transmission line with aperture for simulation.

the Ph.D work. According to equation (3.2), the characteristic impedance of the transmission line shown in Fig. 3.2, whose two reference ports' impedances are both 50Ω , can be obtained by this simulation, the conversion between S-parameters and ABCD-parameters [117] and the following equations.

$$Z_0 = \sqrt{\frac{B}{C}} \quad (3.2)$$

$$\cosh(\beta l) = \frac{A + D}{2} \quad (3.3)$$

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$$\epsilon_{re} = \beta^2 / k^2 \quad (3.4)$$

The phase constant β or the effective dielectric constant ϵ_{re} of the fixed-length uniform transmission line can be calculated according to the equation (3.3) and (3.4), where k is the wave number in free space [1].

Fig. 3.3 Calculated results of the microstrip line with different aperture width W_a and strip width W_s (a) Characteristic impedance (b) Phase constant

As can be seen from Fig. 3.3 (a), the implementations of adding backed-aperture and narrowing strip widths are feasible in the impedance enhancement. When microstrip width W_s keeps unchanged, the characteristic impedance Z_0 increases as the aperture width W_a varies from 0 to 3.0mm. When the aperture width W_a is fixed, Z_0 moves up as strip width W_s is reduced from 3.0 to 0.2mm. In particular, Z_0 is significantly raised from about 18Ω to over 100Ω at two extreme cases. The increment of Z_0 accompanies with the decrement of the effective dielectric constant ϵ_{re} as listed in Table 3.1.

Table 3.1. Effective dielectric constant of different microstrip lines

Ws	Wa	0	0.6mm	1.0mm	1.6mm	2.0mm	2.6mm	3.0mm
	ϵ_{re}							
0.2mm		6.73	6.58	6.27	5.94	5.77	5.6	5.5
0.5mm		7.18	6.86	6.56	5.97	5.79	5.63	5.58
1.0mm		7.64	7.27	6.83	6.14	5.87	5.63	5.58
3.0mm		8.77	8.56	8.28	7.76	7.42	6.9	6.58

3.3 SIR Simulation Results

In this section, the characteristics of the SIR with low-high-low impedance configuration obtained by the author of the thesis are discussed. This kind of structure is notable for the shortened length and spurious harmonics suppressions. As has stated in the previous sections, the greater impedance ratio of the two stripline sections, the greater value of the ratio of first harmonics f_1 over the fundamental resonant frequency f_0 , i.e., f_1/f_0 . Therefore, the main task of raising the value of the high impedance section and suppress the low impedance section has been carried out. In this part, three types of stepped impedance resonators are presented and simulated in the ADS momentum software. In order to

demonstrate its multiple resonances in terms of explicit S-parameters later on, the resonators are linked with the two 50Ω microstrip feeding lines through the capacitive gap with loose coupling. The resonant frequencies: dominant frequency and the higher order mode frequencies are illustrated with respect to the low- and high- impedance ratio and the respective physical/electrical length. The microstrip resonator circuits are constructed on the dielectric substrate of RT/Duriod 6010 (permittivity: $\epsilon_r=10.8$ & thickness: $h=0.635\text{mm}$).

3.3.1 SIR Type A

Fig. 3.4 SIR Type A (a) Layout (b) Resonant frequencies under different impedance ratio K

Fig. 3.4 (a) depicts the layout of the first type of stepped impedance resonator, in which the SIR is composed of wide strips at the two sides and a central narrow strip. According to the previous analysis about the microstrip line, widening strip will decrease the characteristic impedance and narrowing down strip increases it. The resonant frequencies are drawn in accordance with different impedance ratio K in Fig. 3.4 (b). In this figure, four groups of curves are drawn when the respective lengths of the two sections l_1 and l_2 are both fixed to 2.4mm. The width of the wider strip W_1 is varied from 1.0mm to 4.0mm while the narrow strip width W_2 keeps unchanged with 0.6mm. As the impedance ratio $K=Z_2/Z_1$ decreases, the ratio of first harmonic f_1 over dominant frequency f_0 i.e. $q_1=f_1/f_0$ increases from 2.16, 2.59, 2.95 to 3.28 while the electrical lengths of the two parts also change a little because of the bended shape of SIR.

Table 3.2 Frequency ratio $q_1=f_1/f_0$, $q_2=f_2/f_0$ versus physical length ratio: SIR Type A

l_2/l	10.4%	19.8%	29.2%	39.6%	50%	58.3%	70.8%	81.3%	89.6%
f_0 (GHz)	3.69	3.38	3.20	3.09	3.06	3.25	3.22	3.44	3.73
f_1 (GHz)	8.12	8.20	8.57	9.12	10.03	10.67	10.91	10.42	9.93
f_2 (GHz)	13.0	13.70	14.62	15.64	15.42	14.54	13.10	12.24	11.91
f_1/f_0	2.20	2.43	2.68	2.98	3.28	3.29	3.39	3.03	2.66
f_2/f_0	3.52	4.05	4.57	5.06	5.04	4.47	4.07	3.56	3.19

Actually the impedance ratio is not the only factor that makes influence on the frequency ratio q_1 , q_2 and thus making the harmonics far away from or come close to the dominant frequency. It is tabulated in Table 3.2 to show how the physical length affects this parameter as well in the condition of $W_1=4.0\text{mm}$. We can see that frequency ratio q_1 has maximum point around the point of the length percentage of the low-impedance section l_2 over whole length ($l=l_1+l_2$) at about 58.3%-70%, $q_2=f_2/f_0$ has maximum around 39.6%.

3.3.2 SIR Type B

The second type of the SIR is a uniform microstrip line with a square aperture at the center and its layout is described in Fig. 3.5 (a). Unlike the previous type A, the strip

Fig. 3.5 SIR Type B (a) Layout (b) Resonant frequency under different impedance ratio K

width W_2 is fixed to be 0.6mm but the central square aperture width W_1 is chosen to be 1.0mm, 2.0mm, 3.0mm and 4.0mm in the resonant frequency comparison in Fig. 3.5 (b). In this figure, $q_1=f_1/f_0$ increases from 2.07 to 2.80 at the smallest and biggest aperture widths, which is consistent with previous results. Like Table 3.2, Table 3.3 gives the

relation between the physical length and the frequency ratio at $W_l=4.0\text{mm}$. In the table, $q_1=f_1/f_0$ has maximum value around 58.3% and $q_2=f_2/f_0$ at about 39.6%.

Table 3.3 Frequency ratio $q_1=f_1/f_0$, $q_2=f_2/f_0$ versus physical length ratio: SIR Type B

l_2/l	10.4%	19.8%	29.2%	39.6%	50%	58.3%	70.8%	81.3%	89.6%
$f_0(\text{GHz})$	5.27	4.95	4.73	4.59	4.52	4.52	4.59	4.73	4.90
$f_1(\text{GHz})$	10.93	10.99	11.31	11.87	12.45	12.65	12.22	11.57	11.10
$f_2(\text{GHz})$	16.90	17.67	18.62	18.80	17.30	16.0	14.53	13.75	13.43
f_1/f_0	2.07	2.22	2.39	2.59	2.75	2.80	2.66	2.45	2.26
f_2/f_0	3.21	3.57	3.94	4.10	3.83	3.54	3.17	2.91	2.74

3.3.3 SIR Type C

(a)

(b)

Fig. 3.6 SIR Type C (a) Layout (b) Resonant frequency under different impedance ratio K

In the configuration of the SIR Type C structure, the previous methods are combined together. In Fig. 3.6 (a), this SIR is composed of a central narrow strip with backed-aperture and two broadened strips at sides. Similar with the analysis of SIR B and C, the curves of resonant frequencies versus different impedance ratio are depicted in Fig. 3.6 (b) and related parameters listed in Table 3.4. Due to the great increment of K in Type C, the average ratio q_1 is much larger than that in Type A and B. Also, it is observed that q_1 reaches its maximum point around 58.3% and q_2 around 39.6% concerning the physical length ratio l_2/l . According to the above simulated frequency response of the three types of SIR, the frequency ratio is changed with the electrical/ physical length and characteristic impedance of the two strip sections. Due to the bent shape of the wider strip in SIR, the respective electrical length cannot be accurately computed though the phase constant is approximated. The fundamental frequency f_0 is smallest when $l_2/l \cong 50\%$ in all three cases. It means that the resonator has a shortest length under this condition when f_0 is specified. In the selection of the electrical length, there is a tradeoff between the normalized resonant length and the frequency ratio. It is found that adding the aperture does not have significant influence on the simulated insertion loss.

Table 3.4 Frequency ratio $q_1=f_1/f_0$, $q_2=f_2/f_0$ versus physical length ratio: SIR Type C

l_2/l	10.4%	19.8%	29.2%	39.6%	50%	58.3%	70.8%	81.3%	89.6%
$f_0(\text{GHz})$	2.83	2.49	2.31	2.09	2.18	2.20	2.35	2.54	2.88
$f_1(\text{GHz})$	8.59	9.08	9.86	11.03	12.44	13.04	11.95	10.82	10.02
$f_2(\text{GHz})$	15.37	15.65	16.20	18.70	16.45	14.70	12.65	11.50	10.91
f_1/f_0	3.03	3.65	4.27	5.28	5.71	5.93	5.09	4.26	3.48
f_2/f_0	5.43	6.29	7.01	8.95	7.54	6.68	5.38	4.53	3.78

3.3.4 Q-factors of the Resonators

Fig. 3.7 Resonant frequency (f_0) and Q-factor of the microstrip line SIR in Fig. 3.6 (a) ($l_2=2l_1=2$ mm). (a) versus strip width (W_2) in the middle without aperture (b) versus aperture width under the fixed $W_2=0.2$ mm.

The unloaded Q-factor (Q_0) and resonant frequency (f_0) of the SIR resonator in Fig. 3.6 (a) are further extracted from fullwave simulation, aiming to comparatively investigate the loss behaviors of the considered SIR resonator. These two parameters versus strip and aperture widths are plotted in Fig. 3.7 (a) and 3.7 (b), respectively, in conjunction with the two SIRs without and with backed aperture. From Fig. 3.7 (a), we can see that as W_2 is narrowed, resonant frequency f_0 gradually drops off while Q_0 is quickly increased at first and then decreases beyond the peak of about 150. This non-monotonic variation is mainly caused by opposite frequency-varied tendency of radiation and conductor losses or unloaded Q factors. Then, the aperture is formed underneath the narrow strip with $W_2=0.2$ mm. Fig. 3.7(b) illustrates that as the aperture width (W_a) is widened, Q_0 is raised to 126 from 116 (case of no aperture). Meanwhile, f_0 can be further reduced with W_a . As such, such a backed-aperture in the middle section can not only minimize the overall length of this SIR, but also increase its unloaded Q-factor.

3.4 Experimental Verification

Three resonators: UIR, SIR (1) and (2) were designed, simulated and fabricated to make a comparison of measured results with the predicted ones. Although the strip width can be further widened or narrowed in theory and simulation, the SIR performance may suffer from the tolerance of fabrication process, conductor loss and parasitic effect of conductor thickness. Therefore these aspects in my examined frequency range have been taken into consideration in the optimized SIR.

Fig. 3.8 Three resonators for measurement comparison
(a) Layouts (b) Fabricated photograph top view (c) Fabricated photograph bottom view

3.4.1 Layouts of the Fabricated Resonators

The layout of the traditional microstrip uniform-impedance resonator is shown in the top of Fig. 3.8 (a). This UIR is linked with the two 50Ω microstrip feeding lines through the capacitive coupling gap as well. The second and third one in Fig. 3.8 (a) illustrate the

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considered microstrip SIR resonator (1) and (2), in which (1) is like the SIR type A and (2) is composed of the aperture-backed microstrip section with narrow strip width at center and the two identical ones with wide strip width at two sides. The former and latter sections are constructed in order to realize the high and low characteristic impedances, Z_1 & Z_2 , respectively. As studied in [116] and microstrip line characteristics stated above, this backed aperture underneath the narrow strip allows one to reduce the surface-to-surface capacitive coupling between the strip and ground plane to a great extent as requested in realizing the high impedance (Z_1). As the impedance ratio, Z_2/Z_1 , of the two segments in the SIR increases, the 1st spurious resonant frequency is pushed up away from its corresponding fundamental one (f_0) [34], resulting in widening the upper rejection band.

Before the experiment, filtering performances of the microstrip UIR and SIR circuits were simulated using the Agilent Momentum software. For the comparison, the overall lengths of these three resonators were optimized towards the identical fundamental resonant frequency of $f_0=4\text{GHz}$. After careful adjustment in simulation, it was found that they were in the length of 13.4mm, 8.4mm and 6.0mm in conjunction with the UIR and SIR resonators in Fig. 3.8 (a), implying that the proposed two SIR has achieved the size reduction by 37% and 55% against the traditional UIR resonator respectively.

3.4.2 Simulated and Measured Results

Fig. 3.9 (a) shows the simulated magnitude of S_{21} for the microstrip UIR and SIR circuits with the dimensions as indicated in Fig. 3.8 (a). For the UIR, the 1st, 2nd, 3rd and 4th spurious resonances occur at the approximately integral multiples of its fundamental resonant frequency (f_0). Of them, the 1st harmonic frequency (f_1) is observed at 7.8GHz, so

Fig. 3.9 S-parameters $|S_{21}|$ (a) Simulated results (b) Measured results

the aspect ratio of the first two frequencies is deviated as $f_1/f_0=1.9$. For the proposed SIR, the 1st spurious frequency (f_1) is dramatically increased to about 12.3GHz in (1) while 18.3GHz in (2) beyond the 1st, 2nd and 3rd harmonics of the traditional UIR resonator. The aspect ratio now becomes $f_1/f_0=3.08$ and 4.58, implying that this proposed SIR resonator has achieved an ultra-broad rejection band between f_0 and f_1 . The above three resonator circuits are then fabricated as the photographs of the fabricated circuits are shown in Fig. 3.8 (b) and (c). Three sets of measured S_{21} -parameters are plotted in Fig. 3.9 (b), indicating a good verification on our predicted results of Fig. 3.9 (a). Again, these aperture-backed SIR resonator evidently show the actual realization of an ultra-broad stopband in experiment.

3.5 Conclusion

As the characteristics impedance varies with the microstrip width and is affected by the aperture on the ground, an improved microstrip SIR resonator is proposed in this chapter by forming the backed aperture underneath the central narrow strip section. This novel SIR with enhanced impedance ratio of the low and high impedance sections not only miniaturizes its overall size, but also completely eliminates the first three spurious resonances existing in the traditional UIR. The simulated and measured results demonstrate that this SIR has achieved an ultra-broad rejection bandwidth, ranging from 4.0 to 18.3GHz.

Chapter 4

Bandpass Filters Using SIR

4.1 Introduction

The proposal of a stepped impedance resonator [34] has made contributions in exploring a number of advanced microstrip bandpass filters with widened upper stopband and miniaturized size in the past several years, e.g., [42], [119]. To further increase the ratio of the 1st harmonic to the dominant resonant frequency (f_1/f_0) in SIR, an improved SIR with a backside aperture in the central part, already stated in previous chapter, was proposed in [119] on a basis of the so-called aperture compensation technique [110]. This aperture-backed SIR with an extremely large ratio of f_1/f_0 supposes to be used in this work as a

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basic resonator for design of multi-pole bandpass filters with ultra-broad stopband. On the other hand, the microstrip open-end gap or its modified coupling structure is found to be usually suffered from very weak coupling extent that is not sufficiently large as requested in the design of bandpass filters with a relatively wide fractional bandwidth and low insertion loss, especially at low frequencies [44]. In addition to the existing tolerance in fabrication process, the minimum strip or gap width must be at least a few mils so as to minimize the conductor loss and parasitic effects of strip thickness. Following the recent work in CPW filter design [51], the microstrip line quarter-wavelength transformers with low and high impedances are investigated in my work as two alternative capacitive and inductive coupling elements or J- and K-inverter elements with enhanced degree. In particular, an extra aperture is formed on ground plane underneath a narrow strip conductor, building up an impedance transformer with large-extent capacitive coupling degree.

In the following sections of this chapter, coupling behaviors of microstrip capacitive gap and two microstrip impedance transformers are firstly characterized in terms of equivalent J- and K-inverter parameters. With the use of the aperture-backed SIR and these coupling elements, two three-pole bandpass filters are proposed, and designed on the RT/Duroid 6010 substrate with the permittivity of $\epsilon_r=10.8$ and thickness of $h=0.635\text{mm}$. Predicted filtering performances are finally confirmed in experiment, exhibiting an ultra-broad rejection band beyond the dominant passband.

4.2 Coupling Structures

Fig. 4.1 K-impedance inverter (a) Inductance T-network (b) K-inverter network

The basic schematics and characteristics of the K- and J-inverters are discussed in this part. The conversions or inter-relationship between the S-parameters, Y- and Z-matrix parameters and transmission ABCD matrix [117] are implemented in the J- and K-inverter equivalent circuit model analysis.

Fig. 4.2 Equivalent circuit model of K-inverter (a) Z-matrix (b) Detailed K-inverter T-network

4.2.1 K-inverter Block

Fig. 4.1 (a) depicts the typical T-network of a lumped K-inverter impedance with a positive inductance L in shunt and two series negative ones $-L$, making an impedance inverter with characteristic impedance $K=1/\omega L$. Fig. 4.1 (b) shows its exact K-inverter block. Now, let's consider a general two-port Z-matrix block in Fig. 4.2 (a) with asymmetrical geometry with respect to its two ports. Fig. 4.2 (b) shows its resulting K-inverter impedance with two unequal electrical lengths, $\Phi_1/2$ and $\Phi_2/2$, in two sides.

The ABCD matrix in Fig. 4.2 (a) is expressed in terms of Z-parameters as [1], [117]:

$$\begin{bmatrix} A_z & B_z \\ C_z & D_z \end{bmatrix} = \begin{bmatrix} X_{11}/X_{12} & (X_{11}X_{22} - X_{12}^2)/X_{12} \\ 1/X_{12} & X_{22}/X_{12} \end{bmatrix} \quad (4.1)$$

where $Z_{11}=jX_{11}$, $Z_{12}=jX_{12}=Z_{21}=jX_{21}$ and $Z_{22}=jX_{22}$ for the lossless case.

The ABCD matrix of the K-inverter impedance itself can be simplified as:

$$\begin{bmatrix} 1 & -jK \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ -j\frac{1}{K} & 1 \end{bmatrix} \begin{bmatrix} 1 & -jK \\ 0 & 1 \end{bmatrix} = \begin{bmatrix} 0 & -jK \\ -j\frac{1}{K} & 0 \end{bmatrix} \quad (4.2)$$

The total network in Fig. 4.2 (b) is composed of the three-cascaded segments and its

ABCD matrix can be expressed as multiplication of three sub-matrices:

$$\begin{bmatrix} A_K & B_K \\ C_K & D_K \end{bmatrix} = \begin{bmatrix} \cos \Phi_1/2 & jZ_{01} \sin \Phi_1/2 \\ jY_{01} \sin \Phi_1/2 & \cos \Phi_1/2 \end{bmatrix} \begin{bmatrix} 0 & -jK \\ \frac{1}{jK} & 0 \end{bmatrix} \begin{bmatrix} \cos \Phi_2/2 & jZ_{02} \sin \Phi_2/2 \\ jY_{02} \sin \Phi_2/2 & \cos \Phi_2/2 \end{bmatrix} \quad (4.3)$$

Under the exact equivalence of equation (4.1) and (4.3), K , Φ_1 and Φ_2 can be obtained in terms of Z-parameters:

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$$\Phi_1 = \arctan \frac{2(\bar{X}_{11} + \bar{X}_{22} |\bar{X}|)}{1 + \bar{X}_{22}^2 - \bar{X}_{11}^2 - |\bar{X}|^2} + n\pi \quad (4.4)$$

$$\Phi_2 = \arctan \frac{2(\bar{X}_{22} + \bar{X}_{11} |\bar{X}|)}{1 + \bar{X}_{11}^2 - \bar{X}_{22}^2 - |\bar{X}|^2} + n\pi \quad (n=0,1,2,3\dots) \quad (4.5)$$

$$\bar{K} = \frac{K}{\sqrt{Z_{01}Z_{02}}} \frac{-\sin(\Phi_1/2) + \bar{X}_{11} \cos(\Phi_1/2)}{\bar{X}_{12} \sin(\Phi_2/2)} = \frac{\bar{X}_{12} \cos(\Phi_2/2)}{\bar{X}_{11} \sin(\Phi_1/2) + \cos(\Phi_1/2)} \quad (4.6)$$

where $\bar{X}_{11} = X_{11}/Z_{01}$, $\bar{X}_{12} = X_{12}/\sqrt{Z_{01}Z_{02}}$, $\bar{X}_{22} = X_{22}/Z_{02}$ are the normalized reactance with $|\bar{X}| = \bar{X}_{11}\bar{X}_{22} - \bar{X}_{12}^2$.

4.2.2 J-inverter Block

Fig. 4.3 J-admittance inverter (a) Capacitance π -network (b) J-inverter network

Regarding the lumped J-inverter admittance, the typical π -network in Fig. 4.3 (a) has a series positive capacitance C and two shunt negative capacitance $-C$. its characteristic admittance can be expressed as $J=1/\omega C$. the negative C may be absorbed into adjacent series or parallel resonators so as to eliminate them from the overall filter network in

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practical design. Fig. 4.3 (b) shows its exact J-inverter block. Let's consider a general two-port Y-matrix block in Fig. 4.4 (a) with asymmetrical geometry with respect to its two ports. Fig. 4.4 (b) shows the J-inverter admittance with two unequal electrical lengths, $\Phi_1/2$ and $\Phi_2/2$, in two sides.

The ABCD matrix in Fig. 4.4 (a) is formulated in (4.7) in terms of Y-parameters:

$$\begin{bmatrix} A_Y & B_Y \\ C_Y & D_Y \end{bmatrix} = \begin{bmatrix} -B_{22}/B_{12} & -1/B_{12} \\ (B_{12}^2 - B_{11}B_{22})/Y_{12} & -B_{11}/B_{12} \end{bmatrix} \quad (4.7)$$

Where $Y_{11}=jB_{11}$, $Y_{12}=jB_{12}$, $Y_{21}=jB_{21}$ and $Y_{22}=jB_{22}$.

Fig. 4.4 Equivalent circuit model of J-inverter (a) Y-matrix (b) J-inverter network

The ABCD matrix of the J-inverter itself is illustrated as:

$$\begin{bmatrix} 1 & 0 \\ -jJ & 1 \end{bmatrix} \begin{bmatrix} 1 & -j\frac{1}{J} \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ -jJ & 1 \end{bmatrix} = \begin{bmatrix} 0 & -j\frac{1}{J} \\ -jJ & 0 \end{bmatrix} \quad (4.8)$$

Therefore the ABCD matrix of the three-segment J-network in Fig. 4.4 (b) should be

$$\begin{bmatrix} A_J & B_J \\ C_J & D_J \end{bmatrix} = \begin{bmatrix} \cos \Phi_1 / 2 & jZ_{01} \sin \Phi_1 / 2 \\ jY_{01} \sin \Phi_1 / 2 & \cos \Phi_1 / 2 \end{bmatrix} \begin{bmatrix} 0 & \frac{1}{jJ} \\ -jJ & 0 \end{bmatrix} \begin{bmatrix} \cos \Phi_2 / 2 & jZ_{02} \sin \Phi_2 / 2 \\ jY_{02} \sin \Phi_2 / 2 & \cos \Phi_2 / 2 \end{bmatrix} \quad (4.9)$$

Thus the normalized J-inverter, phase Φ_1 and Φ_2 can be obtained by equalizing equation (4.7) to (4.9) in the form of the Y-parameters:

$$\Phi_1 = \arctan \frac{2(\bar{B}_{11} + \bar{B}_{22} | \bar{B} |)}{1 + \bar{B}_{22}^2 - \bar{B}_{11}^2 - | \bar{B} |^2} + n\pi \quad (4.10)$$

$$\Phi_2 = \arctan \frac{2(\bar{B}_{22} + \bar{B}_{11} | \bar{B} |)}{1 + \bar{B}_{11}^2 - \bar{B}_{22}^2 - | \bar{B} |^2} + n\pi \quad (n=0,1,2,\dots) \quad (4.11)$$

$$\bar{J} = \frac{J}{\sqrt{Y_{01}Y_{02}}} = \frac{-\bar{B}_{12} \cos(\Phi_2 / 2)}{\bar{B}_{11} \sin(\Phi_1 / 2) + \cos(\Phi_1 / 2)} = \frac{-\bar{B}_{11} \cos(\Phi_1 / 2) + \sin(\Phi_1 / 2)}{\bar{B}_{12} \sin(\Phi_2 / 2)} \quad (4.12)$$

where $\bar{B}_{11} = B_{11} / Y_{01}$, $\bar{B}_{12} = B_{12} / \sqrt{Y_{01}Y_{02}}$, $\bar{B}_{22} = B_{22} / Y_{02}$ are the normalized reactance

with $| \bar{B} | = \bar{B}_{11} \bar{B}_{22} - \bar{B}_{12}^2$.

4.3 Capacitive Gap Coupling

The microstrip gap coupling element is the most commonly used capacitive coupling structure in microstrip filters and its geometry is illustrated in Fig. 4.5. This structure is

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formed by closely spacing the two open-ends of two microstrip lines with width W and characteristic impedance Z_{01} . In this aspect, the S-parameter of this two-port circuit is at first calculated via the Agilent Momentum software. Then, based on the equations (4.10)-(4.12), the equivalent J-inverter and electrical length Φ_1 and Φ_2 can be derived. Due to the symmetrical geometry, Φ_1 must be equal to Φ_2 . Fig. 4.6 depicts the calculated mutual- and self-conductance and susceptance of the Y-network at the fixed reference line width $W=0.1\text{mm}$ and different gap S . In these two figures, it is shown that the self-susceptance B_{11} and mutual-susceptance B_{12} are much greater than the self-conductance G_{11} and mutual-conductance G_{12} . Accordingly, the loss-related conductance part can be fully neglected in the filter design.

Fig. 4.5 Layout of the microstrip capacitive gap coupling

Fig. 4.6 Calculated G_{11} , G_{12} , B_{11} , and B_{12} of the capacitive gap under different gap width
(a) G_{11} and G_{12} (b) B_{11} and B_{12}

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(a)

(b)

Fig. 4.7 Calculated coupling parameters of capacitive gap
(a) Normalized J-inverter susceptance (b) Phase $\Phi_1=\Phi_2$

Fig. 4.7 is the calculated results of the normalized J-inverter and electrical length of the microstrip gap under different dimensions. The J-inverter or the coupling degree of this gap increases linearly with the frequency. At the same width of $W=1.0\text{mm}$ (three low lines), we can find that the smaller gap distances results in larger couplings. Under the identical gap distance 0.1mm (upper two lines), the thinner strip presents a small coupling. The two electrical lengths, $\Phi_1=\Phi_2$, are non-linearly with frequency.

4.4 $\lambda/4$ Impedance Transformers

Fig. 4.8 (a) and (b) depict the layouts of the two microstrip line quarter-wavelength transformers with high and low impedances or narrow and wide strip widths, respectively. In the analysis, both of them can be considered as a unified microstrip line section, with the width (W_2) and characteristic impedance (Z_{02}), that is driven at the two sides by the feed lines with the strip width (W_1) and characteristic impedance (Z_{01}). When $W_1=0.6\text{mm}$, $Z_{01}=50\Omega$ with substrate thickness $h=0.635\text{mm}$ and $\epsilon_r=10.8$. As pointed out in [51], they can be perceived as equivalent J- and K-inverter networks as indicated in Fig. 4.8 (c) and (d). To initially show their coupling performances at the frequency of one quarter-wavelength, their relevant normalized J-inverter admittance and K-inverter impedance of these two coupling structures are calculated and plotted in Fig. 4.8 (e) with respect to impedance ratio (Z_{01}/Z_{02}).

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Fig. 4.8 The impedance transformers (a) High impedance transformer layout (b) Low impedance transformer layout (c) J-inverter circuit model (d) K-inverter circuit model (e) Normalized J and K inverter at the $\lambda/4$ frequency.

4.4.1 Low-Impedance Transformer

Fig. 4.9 Equivalent circuit models of the low-impedance transformer: Transmission line
To realize the coupling network in Fig. 4.8 (d), the transformer with wide strip width ($W_2 > W_1$) in Fig. 4.8 (b) is at first perceived as an equivalent uniform transmission line section with small impedance ($Z_{02} < Z_{01}$) as shown in Fig.4.9. Then, its relevant ABCD matrix can be expressed as [1]:

$$\begin{bmatrix} A_T & B_T \\ C_T & D_T \end{bmatrix} = \begin{bmatrix} \cos(\beta_2 l_2) & jZ_{02} \sin(\beta_2 l_2) \\ jY_{02} \sin(\beta_2 l_2) & \cos(\beta_2 l_2) \end{bmatrix} \quad (4.13)$$

Next, this initial network is converted to a coupling-related network with K-impedance and the two identical transmission line lengths ($\Phi_1/2 = \Phi_2/2$) at the two sides as shown in Fig. 4.2 (b). By enforcing the equalization of Equations (4.13) and (4.3), the explicit relationships among K, Φ_1 , Φ_2 and Z-parameters can be obtained as done in [48].

$$K / Z_{01} = \frac{\cos(\beta_2 l_2)}{\sin \Phi_1} - \frac{1}{2} \left(K_z - \frac{1}{K_z} \right) \sin(\beta_2 l_2) \quad (4.14)$$

$$\Phi_1 = \Phi_2 = -\arctan \left(\frac{2 \cos(\beta_2 l_2)}{(K_z + 1/K_z) \sin(\beta_2 l_2)} \right) \quad (4.15)$$

where K_z is the impedance ratio $K_z = Z_{02}/Z_{01}$

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Equations (4.14) and (4.15) can be utilized to evaluate quantitatively the inductive coupling extent of a low-impedance transformer in terms of explicit K-network parameters. Fig. 4.10 (a) and (b) plot the normalized K-impedance (K/Z_{01}) and two equal phases ($\Phi_1/2$) over the frequency range of 2.0 to 6.0 GHz under the three different strip widths (W_2).

Fig. 4.10 Simulated equivalent K-network parameters
(a) Normalized K-inverter impedance (b) Phase ($\Phi_1/2=\Phi_2/2$)

As compared to $K/Z_{01} < 0.2$ for the open-end gap in [44] and previous section, this quantity becomes 1.0 (unity) at the beginning when $W_2 = W_1 = 0.6\text{mm}$ and gets a gradual decrease to 0.74, 0.51 and 0.35 as W_2 is widened to 1.0, 2.0 and 3.0mm. Looking at Fig. 4.10 (a) and (b) together, we can further find that the equal phases $\Phi_1/2$ at two sides reaches zero degree at the frequency close to $f_0 = 4\text{GHz}$.

4.4.2 High-Impedance Transformer

Similarly, in order to study capacitive coupling behaviors of the high-impedance transformer in Fig. 4.8 (a), the two relevant network topologies are illustrated in Fig. 4.11 and Fig. 4.4 (b). The transmission line model in Fig. 4.11 with high impedance ($Z_{02} > Z_{01}$) still can be characterized in terms of Equation (4.13). The ABCD matrix of J-network in Fig. 4.4 (b) can be obtained by multiplying the three-cascaded sub-matrices as done before in Equation (4.8)

Fig. 4.11 Equivalent circuit models of the high-impedance transformer: Transmission line

Under the exact equivalence of the two networks in Fig. 4.11 and 4.4 (b), Equations (4.13) and (4.9) must be equalized with each other so as to allow one deriving normalized J-

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inverter (J/Y_{01}) and two same phases ($\Phi_1=\Phi_2$) in the explicit format [26].

$$J/Y_{01} = \frac{\cos(\beta_2 l_2)}{\sin \Phi_2} + \frac{1}{2} \left(K_z - \frac{1}{K_z} \right) \sin(\beta_2 l_2) \quad (4.16)$$

$$\Phi_1 = \Phi_2 = -\arctan \left(\frac{2 \cos(\beta_2 l_2)}{(K_z + 1/K_z) \sin(\beta_2 l_2)} \right) \quad (4.17)$$

Fig. 4.12 Equivalent J-network parameters
(a) Normalized J-inverter (b) Phase ($\Phi_2/2=\Phi_1/2$)

Fig. 4.12 (a) and (b) indicate the simulated results using Equations (4.16) and (4.17). Normalized J-inverter admittance (J/Y_{01}) tends to gradually fall down and eventually reach to 0.55 as W_2 is reduced from 0.6 to 0.4, 0.2 and 0.1mm. This coupling quantity is minimized when Φ_2 approaches zero degree near 4.0GHz as indicated in Fig. 4.12 (b).

4.4.3 Aperture-Backed Impedance Transformer

The above-derived J/Y_{01} with the minimum value of 0.55 at $W_2=0.1$ mm is found still too large in the design of a microstrip bandpass filter with the fractional bandwidth (FBW)

Fig.4.13 Aperture-backed microstrip line high-impedance transformer with the fixed $W_2=0.2$ mm. (a) Layout (b) Calculated normalized J/Y_{01} .

larger than 5.0%. In order to further reduce this coupling quantity, the strip width (W_2) is required to become narrower and narrower. However, it may suffer from several unexpected problems as discussed in [110], [119], such as extra conductor loss and difficult-to-estimated parasitic effects of conductor thickness. Using the aperture compensation technique [110], this problem may be overcome by forming a backside aperture underneath a relatively narrow strip conductor as shown in Fig. 4.13(a). Fig. 4.13 (b) shows the normalized J-inverter admittance (J/Y_{01}) versus aperture width (W_a) under the fixed strip width ($W_2=0.2\text{mm}$). We can see here that J/Y_{01} is reduced below 0.5 and it can become further small if W_a is widened.

4.5 SIR Bandpass Filters

In this section, the above-discussed microstrip high and low impedance transformer with tightened J- or K- inverter and the multistage SIR with backed-aperture are utilized to construct the two novel types of bandpass filters. They are compared with the first type of bandpass filters using capacitive gap external coupling. Their attractive ultra-broad stopband behavior is examined in both theory and experiment with its two-port scattering parameters simulated by the fullwave software ADS momentum and measured by the HP network analyzer and test fixture.

4.5.1 SIR Filter Type A

The loose external coupling is used in this part to design the first type of three stage bandpass filter as depicted in Fig. 4.14 (a). The backed-aperture stepped impedance resonators, which are cascaded and end-coupled to each other, have the same dimensions with previously proposed SIR and it has the first dominant resonant frequency to be fixed around 4.0GHz. All the couplings are capacitively end-coupled by gaps. The inter-resonator coupling should be much weaker than that between the feeders and the first/last stage resonator according to the filter design theorem.

Fig. 4.14 BPF Type A (a) Layout ($\Delta_1=0.005\text{mm}$, $\Delta_2=0.1\text{mm}$, $g_2=0.545\text{mm}$)
(b) S_{11} -magnitudes under different external couplings.

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As analyzed in section 4.3, the capacitive gap, characterized as J-inverter with two electrical lengths $\Phi_1/2$ and $\Phi_2/2$, provides couplings and the electrical lengths are absorbed by the adjacent resonators. During the optimization of the bandpass filters, the center SIR should have some dimension change i.e. Δ_1 and Δ_2 . To illustrate the gap coupling effects, Fig. 4.14 (b) plots the simulated S_{11} -magnitudes of different external coupling degrees with fixed inter-resonator distance g_2 . In the feed lines, the steps are stretched out to have a large coupling area with the first/last stage SIR. It is observed from Fig. 4.14 (b) that the filter has better passband performance with strengthened external coupling degree. The solid line case has the best S_{11} -magnitude with return loss higher than 10dB and three balanced transmission poles.

Herein, the dominant central frequency of the optimized bandpass filter is designed at 3.91GHz. The simulated S-parameters are drawn in Fig. 4.15 (a) and (b) in a wide or narrow band. In the wide frequency range from 2-20GHz in Fig. 4.15 (a), the 1st harmonic are shifted away from the dominant frequency to 4.86 times of it, appearing at about 19.0GHz and creating a ultra-wide stopband with return loss mostly higher than 40dB between them. When zooming into the narrowband view in Fig. 4.15 (b), we can see that the bandpass filter has a 3.1% fractional bandwidth with insertion loss lower than 1dB and return loss higher than 12dB.

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Fig. 4.15 Simulated S-parameters of the optimized BPF Type A (a) Wide-band (b) Narrow-band

4.5.2 SIR Filter Type B

(a)

(b)

Fig. 4.16 Layout of the three-stage UIR and proposed SIR bandpass filter Type B
 (a) UIR filter layout (b) SIR filter layout dimensions: $L_1=L_2=2.0\text{mm}$, $W_1=0.6\text{mm}$,
 $W_2=0.2\text{mm}$, $W_3=3.0\text{mm}$, $W_4=1.5\text{mm}$, $\Delta=-0.01\text{mm}$, $g=0.1\text{mm}$

The fractional bandwidth is relatively small in the above BPF Type A because of the insufficient external coupling degree. Now that the external coupling should be several magnitudes larger than the inter-resonator coupling, the inter-resonator gap g is set to smallest tolerable distance 0.1mm for the sake of fabrication. The gap cannot solely satisfy the external coupling but the previously analyzed high-impedance transformer, characterized as J- inverter admittance, is able to reach such a strong degree. Fig. 4.16 (a) and (b) illustrate the layouts of two three-stage bandpass filters with aperture-backed high-impedance transformers that are employed at input and output ports as enhanced

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capacitive coupling elements. For a comparison study, the former one uses the traditional uniform impedance resonators with one half-wavelength in length, whereas the latter one employs the aperture-backed SIR resonator as discussed above. In both filters, the opened-end gap with relatively weak coupling is introduced between two of adjacent SIR resonators as shown in Fig. 4.16.

Fig. 4.17 Simulated S-parameters (a) S-parameters magnitudes under different external couplings in passband (b) Optimized S_{21} -magnitudes comparison of SIR and UIR BPF Type B in wideband

Fig. 4.17 (a) depicts the S-parameters under different external couplings with other dimensions of the SIR filter in Fig. 4.16 (b) unchanged. When the filter is insufficiently coupled via the aperture width $W_2=2.5\text{mm}$, the return loss is high. But when it is over-coupled at $W_2=0.5\text{mm}$, the return loss does not meet the requirements to generate balance transmission poles due to the inadequate inter-resonator coupling. The optimized S_{21} -magnitude results of the filters using UIR and SIR resonators over the wideband are comparatively presented in Fig. 4.17 (b) according to the dimensions marked in Fig. 4.16.

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The two filters are optimally designed at the same operating frequency of about 4.0GHz. At first, we can find that the overall filter length is largely reduced from 55.2mm (UIR case) to 36.0mm (SIR case). It is mainly brought by size reduction of the proposed SIR resonator. This SIR bandpass filter has a wider bandwidth 9.9% from 3.75-4.14GHz. Herein which the insertion loss does not exceed 0.5dB with return loss higher than 26dB as shown in Fig. 4.17 (a). For the UIR filter, its spurious resonant frequencies appear at nearly multiple integral numbers of the dominant one if the parasitic end effects of each resonator are ignored. As shown in Fig. 4.17(b), the first three ones in the UIR filter are located near 7.8, 11.2 and 14.9GHz, respectively. As the aperture-backed SIR resonator is used in the SIR filter, all the first three spurious or harmonic passbands are removed out, thus achieving an ultra-broadband upper stopband of 4.4GHz to 17.8GHz under the insertion loss higher than 20dB.

Fig. 4.18 Layout of the three-stage UIR and proposed SIR bandpass filter Type B (a) UIR filter layout (b) SIR filter layout dimensions: $L_1=L_2=2.0\text{mm}$, $W_1=0.6\text{mm}$, $W_2=0.2\text{mm}$, $W_3=3.0\text{mm}$, $W'_3=4.4\text{mm}$, $\Delta'=0.04\text{mm}$, $g=0.1\text{mm}$

4.5.3 SIR Filter Type C

In this part, the SIR bandpass filter Type C is constructed by using the low-impedance transformer as the external strengthened coupling. In particular, the $\lambda/4$ resonators are used as the 1st and 3rd stage blocks in this type of BPF, which is different from the filter Type B configuration. Fig. 4.18 (a) and (b) depict the layouts of the UIR and SIR bandpass filters that use the low-impedance transformer as a tightened inductive-coupling element. They are still designed at 4.0GHz. In this case, the overall lengths of these two designed filters become 36.9mm and 22.3mm, respectively, and they are much shorter than their counterpart filters shown in Fig. 4.16. The inter-resonator gaps (g) are also set to about 0.1mm for tight coupling. The coupling degree of the low-impedance transformer must be stronger to match the filter design parameters thus having a proper passband characteristic. The transformer width W_3' changes from 3.0m, 4.4m, to 5.0mm, the filtering performances are changed in terms of insufficient-, proper- and over-coupling extent, respectively, as illustrated in Fig. 4.19 (a). At the optimized dimensions of Fig. 4.18, the bandwidth is 13.1% or 3.7GHz to 4.22GHz and return loss is larger than 17dB and insertion loss smaller than 1.4dB. Fig. 4.19 (b) plots the simulated S_{21} -magnitudes of the two designed bandpass filters. Again, the filter with the proposed aperture-backed SIR resonator exhibits complete suppression of the first three harmonics that exist in the traditional UIR filter. Under the insertion loss higher than 20dB, the upper stopband covers an ultra-broad frequency range from 4.4GHz to 16.8GHz.

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Fig. 4.19 Simulated S-parameters (a) SIR BPF Type C S-parameters in passband
(b) S_{21} -magnitude of SIR and UIR BPF Type C in wideband

Fig. 4.20 Fabrication photograph and measured S-parameters of proposed SIR filter Type A (a) Top view (b) Bottom view (c) Measured S-parameters magnitudes

4.6 Experimental Verification

In order to verify the simulated results, the optimized filter Type B is fabricated and its top and bottom views are illustrated in Fig. 4.20 (a) and (b). Fig. 4.20 (c) illustrates the predicted and measured magnitudes of the S_{21} and S_{11} in the wide frequency range from 2 to 20GHz. The measured results has verified the simulated results that the 1st spurious resonant frequency is increased beyond 18.0GHz, over 4 times of central frequency $f_0=3.95$ GHz of the dominant passband. The filter exhibits the ultra-broad rejection band with most of the insertion loss lower than 40dB and a relative widened passband of about 10.0%. Within the dominant passband, the predicted and measured return losses are both higher than 16dB, while the measured insertion loss achieves 1.5dB as compared with 0.5dB in simulation. The bandpass filter Type C is then fabricated. Fig. 4.21 (a) and (b) show its top and bottom views. The ultra wide rejection band is ranging from 4.1GHz to over 16.0GHz. Due to the fabrication tolerance and effects of SMA connector used in the measurement, the passband characteristic is not as good as the Type B using test fixture for testing. In addition, this Type C filter has actually shortened the physical lengths of both the $\lambda/4$ SIRs and low-impedance transformers utilized.

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Fig. 4.21 Fabrication photograph and measured S-parameters of proposed SIR filter Type C
(a) Top view (b) Bottom view (c) Measured S-parameters magnitudes

4.7 Conclusion

In this chapter, several novel microstrip bandpass filters with size-miniaturization and ultra-broad rejection band have been proposed. The aperture-backed SIR resonator has been utilized to suppress the high order spurious harmonics and reach a compact circuit size. Three different types of couplings structures: capacitive gap, quarter-wavelength high-impedance and low-impedance transformers are characterized. They are modeled as

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the J-admittance or K-impedance inverters with weak or tight coupling degree for constructing respective three-stage bandpass filters. Due to the inadequate coupling degree, the capacitive gap causes bandwidth relatively small in Filter Type A. Types B and C using the high- or low-impedance transformer have greatly enhanced the external coupling degree, thus having much wider pass bandwidth while keeping the ultra-wide rejection band. Measured S-parameters in wideband and narrowband have verified the simulated results. Moreover, the two sorts of fabricated bandpass filter using SIR have the size reduction of about 35% and 40%, respectively, compared to their UIR counterparts.

Chapter 5

Dual-Mode Bandpass Filters Using SIRR

5.1 Introduction

A diversity of microstrip dual-mode bandpass filters have been widely studied and implemented in the past few decades since the concept of a microstrip ring dual-mode filter was originally proposed in [52]. Usually, it utilizes two orthogonal resonant modes in a one-wavelength ring or loop resonator to construct such a type of filter. As has been stated in previous chapters, the harmonic resonant frequencies of a uniform impedance ring resonator (UIRR) similarly occur at the integral multiples of the fundamental frequency (f_0). To effectively suppress the first a few harmonics, an improved stepped

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impedance ring resonator (SIRR) can be proposed to build up novel microstrip dual-mode filters. In this chapter, the first dual-mode bandpass filter with the slit perturbation has good performance in the passband but does not have a better stopband. This is because that the slit on the wide strip affects its characteristic impedance. Therefore an improved class of microstrip dual-mode filter is proposed by constructing a novel meander-shaped SIRR with square notches as the perturbation elements. Following the works in [34], [53], this proposed SIRR can be expected to have a great capacity in simultaneously achieving the size reduction and harmonic suppression. In order to tighten the line-to-ring coupling, the two open-end feed lines are readily stretched into the central slits of meander lines as implemented in [53], [69]. Furthermore, the two squared notches are etched out on the corners of a symmetrical or unsymmetrical diagonal line. They are utilized here as a pair of inductive perturbations as studied in [66] to excite the two degenerate resonant modes. The two filter blocks are finally designed and fabricated on the RT/Duroid 6010 substrate with the permittivity of $\epsilon_r=10.8$ and thickness of $h=1.27\text{mm}$ with and without transmission zeros. Measured results show that the proposed dual-mode filter with the passband of 6.6% at 1.52GHz achieves the wide stopband from 1.60 to 5.10GHz. Following the work of the first two meandered SIRR, another SIRR is constructed by readily aperture-backing [116] its four high-impedance sections. In this aspect, the 1st harmonic of this backed-aperture SIRR is first confirmed in theory beyond the four times of its fundamental one, i.e., $4f_0$. The dual-mode filter based on this SIRR is then designed at 1.59GHz and the results have shown its ultra-broad stopband of 1.70~5.80GHz and size reduction of 56.0%.

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Fig. 5.1 Uniform ring dual-mode bandpass filter (a) Layout (b) Network topology

5.2 Dual-Mode Resonant Circuit

Now, let's consider the operating principle of a dual-mode ring resonator filter through a simple uniform ring resonator. As illustrated in Fig. 5.1 (a), the two feed lines are arranged by 90° in the space and a pair of perturbation elements (patches) are introduced at the symmetric plane of the input/output ports. In this case, the two degenerate modes [52] in relation to the two perpendicular feed lines are coupled together via two perturbation elements or patches. Fig. 5.1 (b) illustrates the equivalent network topology

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for this dual-mode bandpass filter. The perturbation elements can be equivalently considered as a pair of shunt capacitors C_p at the symmetric plane of the ring topology. Using the ADS momentum software, Fig. 5.2 plots the simulated S_{21} -magnitudes of the uniform ring bandpass filter with and without the perturbation elements, respectively. Without the perturbation elements, the two degenerate modes seem completely isolated or uncoupled with each other as indicated in the dotted line in Fig. 5.2. With the proper arrangement of such perturbation elements, these two modes can be coupled together so as to make up a two-pole bandpass behavior as shown in the solid line in Fig. 5.2.

Fig. 5.2 Calculated S_{21} - magnitude of a dual-mode bandpass filter

5.3 Dual-Mode Filter with Slits

In this section, the dual mode bandpass filter with slits on the four corners is presented. Using the structure of the improved SIRR with backed-apertures, this dual mode bandpass filter is characterized by simulating its scattering parameters.

Fig. 5.3 Improved stepped impedance ring resonator (a) SIRR B (b) SIRR C

5.3.1 Improved Ring Resonator

The improved structures of the stepped impedance ring resonators are proposed in Fig. 5.3 (a) and (b): SIRR Type B and C. Compared to the SIRR Type A in Fig. 2.6 (b), the bent strips at corners are further widened in both types and the backed-apertures are added underneath the narrow strips in Type C. Fig. 5.4 depicts the magnitudes of the scattering parameters S_{21} under the UIRR in Fig. 2.6 (a) and three SIRRs: Type A, B and C. As the corner strips are getting wider and the side central strip narrower or with additional backed-aperture on the ground, the impedance ratio becomes larger. It results in the increment of the ratio of f_1/f_0 . As can be seen from Fig. 5.4 and the layouts, the dimensions of the four ring resonators are adjusted so as to make their fundamental frequency be the same as 1.6GHz, and the first spurious resonant frequencies are $f_I=3.25\text{GHz}$, $f_{IA}=3.94\text{GHz}$, $f_{IB}=5.08\text{GHz}$ and $f_{IC}=5.42\text{GHz}$ respectively, greatly enhancing the ratio f_1/f_0 from 2.03 to 3.39.

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Fig. 5.4 S_{21} -magnitudes of the four types of ring resonators

5.3.2 Proposed Dual-Mode Bandpass Filters

Fig. 5.5 Layout of the dual-mode bandpass filter with slits

(a) Type A: with transmission zeros (b) Type B: without transmission zeros

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Fig. 5.6 Simulated S-parameters of different external coupling length

(a) S_{21} -magnitude (b) S_{11} -magnitude

The above analyzed stepped impedance ring resonator Type C can be applied to the design of the dual-mode bandpass filters. According to the theory of the dual mode filter, the two feeders are reallocated by placing them in perpendicular direction shown in Fig. 5.5 (a) and (b). In order to strengthen the coupling degree, the additional lines of the feeders are stretched along the side of the ring resonator. The backed-aperture enhances both of the characteristic impedance and the external coupling strength. The four slits on the corner are etched out to serve as the perturbation elements, with two identical length of L_2 along the direction of BD and two identical lengths of L_3 along the line of AC. One of the modes in Fig. 5.5 has the current oriented from left to right side while the other is from the upper to lower side. The stretching length L_1 and the difference between the two groups of slit length $\Delta L = L_3 - L_2$ have the effects on the external coupling and inter-mode coupling degree respectively. Fig. 5.6 (a) and (b) are the simulated S-parameters under

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different external coupling length L_1 as the ΔL is fixed to be 1.48mm. Larger L_1 length makes the return loss in the passband higher to a degree and then back to lower level without poles because this external coupling must match to the inter-mode coupling to make a reasonable passband. When ΔL increases, the inter-mode coupling is tightened and the good frequency response occurs at the best case of $\Delta L=1.63\text{mm}$ at the fixed $L_1=3.0\text{mm}$ as shown in Fig. 5.7.

Fig. 5.7 Simulated S-parameters of different inter-mode coupling degree

(a) S_{21} -magnitude (b) S_{11} -magnitude

The optimized case of this dual-mode bandpass filter is simulated and its S-parameters are illustrated in Fig. 5.8 (a) as the narrow band view and Fig. 5.8 (b) as the wide band view. The above analyzed structure only includes the condition of $\Delta L>0$ and the S_{21} -magnitudes have transmission zeros appearing at the two skirts of the passband. When $\Delta L<0$ in the dotted line in Fig. 5.8, the two transmission zeros do not exist. Both of the two filters

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have 9.1% bandwidth centered at 1.53GHz. But the first spurious harmonic occurs at 4.2GHz, only 2.74 times of the fundamental frequency, when broadening the view to wide-band of 1-7GHz. Although the slits make the first resonant frequency lower due to the increment of the total electrical length, the upper rejection band is narrowed down compared to the original SIRR. It is mainly because the introduction of the slit raises the characteristic impedance of the low-impedance and then makes the higher order harmonics close to the first dominant frequency.

(a)

Fig. 5.8 S-parameters of the optimized circuits (a) Narrow-band view (b) Wide-band view

5.4 Meander-Shaped Dual-Mode Filter

5.4.1 Meander-Shaped SIRR

Fig. 5.9 (a) and (b) depict the layouts of the proposed squared meander-shaped stepped impedance ring resonators without or with the backed-aperture, in which the SIRR are also driven by the two paralleled feed lines via weak capacitive coupling gaps. In the proposed SIRR, each of the four identical sides is composed of the high-impedance meander line with the strip width of 0.2mm at the center and the low-impedance strip line with the strip width of 3.0mm at the two sides around the corners of this squared ring resonator. It should be emphasized that the four meander-line strip sections in the center of each arm are placed above the backed-aperture region towards further enlarging their characteristic impedance in Fig. 5.9 (b).

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Fig. 5.9 Layouts and resonance behaviors of the meander-shaped SIRR resonator circuits
 (a) SIRR geometry without aperture (b) SIRR geometry with aperture
 (c) Frequency responses of S_{21} -magnitudes.

According to the theory reported in [34] and my pervious discussions, it can be understood that the stopband between the 1st two resonant frequencies of this SIRR is basically dominated by the impedance and length ratios of these two distinctive line sections. And therefore these SIRRs can be expected to effectively control the stopband between the fundamental and 1st spurious resonant frequencies, i.e., f_0 and f_1 , in terms of the impedance ratio of the narrow and wide strip sections in Fig. 5.9 (c). The electrical

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characteristics of the UIRR in Fig. 2.6 (a) and the proposed SIRR resonator circuits in Fig. 5.9 (a) and (b) are numerically simulated using the Agilent Momentum software. In order to comparatively study their distinctive features, these three circuits based on UIRR and SIRRs are designed to operate at the same fundamental resonant frequency of $f_0=1.57\text{GHz}$ by adjusting the strip-line lengths in the four ring sides. In this way, as can be found in Fig. 2.6 (a) and Fig. 5.9 (a) (b), the overall dimension is significantly reduced from $19.0\text{mm}\times 19.0\text{mm}$ (UIRR case) to $13.4\text{mm}\times 13.4\text{mm}$ (SIRR case without aperture) and $12.8\text{mm}\times 12.8\text{mm}$ (SIRR case with aperture), thus achieving the size reduction of about 29.5%, 32.6% in length and 50%, 55% in area respectively. Simulated frequency responses of S_{21} -magnitude of these three resonators are plotted in Fig. 5.9 (c). For the UIRR, the first four spurious resonances can be seen to almost occur at the integral multiples of its fundamental resonant frequency (f_0). In comparison to $f_1=3.20\text{GHz}$ in the UIRR, the 1st harmonics in the two SIRRs are largely increased to 5.30GHz and 6.43GHz that are beyond the three and four times of f_0 . In other words, the ratio of the two frequencies, f_1/f_0 , is tremendously enlarged from 2.0 to 3.37 and 4.1, implying its attractive capacity in making a dual-mode filter with small size and harmonic suppression. In the next section, this enlarged ratio factor is utilized to implement an improved class of dual-mode ring filters with broadened stopband.

5.4.2 Filters without Backed-Aperture

Fig. 5.10 Layouts of the two SIRR-based dual-mode filters
(a) Type A: with transmission zeros (b) Type B: without transmission zero.

As discussed in [53], [58], [61], [72], the dual-mode ring filter without the backed side aperture is basically constructed by properly introducing the perturbation elements on the symmetrical ring resonator that is fed by the two orthogonal feed lines. These two dual-mode filters based on the SIRR are constructed by etching out a pair of squared notches on the two faced corners along either of the diagonal lines as illustrated in Fig. 5.10 (a) and (b), respectively. The two perpendicular feed lines with open ends are at first simultaneously stretched into the slits of the meander strip sections at center to realize the tightened line-to-ring coupling [53], [69]. The paired notches are then formed along the diagonal line that is perpendicular or parallel to the symmetrical plane of the two feed lines.

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Fig. 5.11 ADS Momentum frequency responses of the dual mode filter with different perturbation size (a) S_{21} -magnitudes (b) S_{11} -magnitudes

Fig. 5.12 ADS Momentum frequency responses of the dual mode filter with different line-to-ring coupling (a) S_{21} -magnitudes (b) S_{11} -magnitudes

In order to examine the notch perturbation element and line-to-ring coupling effect to the filter design, the first filter in Fig. 5.10 (a) is taken into account for the two filters are only different from the notches position. As the simulation results using the ADS momentum

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software are shown in Fig. 5.11 (a) and (b), the insertion loss is getting lower and return loss higher with the occurrence of the transmission poles when the square notches side width W increases under the largest external coupling with coupling gap $S=0.1\text{mm}$. On the other hand, the external coupling increases with the gap width S changing from 0.6 to 0.1mm . The passband is best at $S=0.4\text{mm}$ in Fig. 5.12. But in this case, the insertion loss is still high at 7dB and return loss is small at 7dB due to both of the insufficient external and inter-modes couplings.

Fig. 5.13 Simulated S-parameters of the two optimized dual mode filters
(a) Narrow passband illustration (b) Wide band illustration

Different from the open-end stub as discussed in [53], [54], [58], [72], the notches in the upper-left and lower-right corners actually behave like an equivalent inductive element. As such, following the theorem in [66], we can intuitively understand that the transmission zeros on the lower and upper rejection skirts of the dominant passband are existing in the Type A and not in the Type B. This can be seen in Fig. 5.13 which shows

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the S-parameters of the two dual mode filters with optimized dimensions. Both of the filters have a fractional bandwidth 6.8% centered at 1.60GHz with 3dB insertion loss and return loss higher than 15dB in Fig. 5.13 (a). In the wideband arrangement of the S-parameters, the higher order harmonics happen beyond 5.2GHz. It is found that both of the 1st dominant frequency and 1st harmonic are not changed much compared with the same structure SIRR without the notch perturbation. The introduction of the square notches divides the current route into two paths thus having two raised equivalent inductances in parallel. And therefore the total inductance and the characteristic impedance of the wide strips are both changed little.

5.4.3 Filters with Backed-Aperture

Fig. 5.14 Layouts of the two SIRR-based dual-mode filters with cross-shape backed-aperture (a) Type C: with transmission zeros (b) Type D: without transmission zeros

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As the meander-shaped SIRR tells that adding backed-aperture can further improve its upper rejection band, Fig. 5.14 (a) and (b) show us the proposed dual-mode SIRR filters with this preferred cross-shaped aperture. It not only increases the high-impedance part of the narrow strip but also enhances the line-to-strip coupling strength. These notches with series-inductive loading result in emergence of the two preferred transmission zeros outside the dominant passband [66]. Fig. 5.15 (a) depicts the simulated S-parameters of the designed dual-mode filter. It is seen that filter Type C has the dominant passband of 7.6% at $f_0=1.57\text{GHz}$, insertion loss less than 2.0dB within the passband. In the filter Type D, it has a 7.7% bandwidth centered at 1.55GHz with high return loss and low return loss as well. Both of them have the ultra-broad stopband of 1.70~5.80GHz under the insertion loss above 10.0dB. The upper rejection band is widened compared to the Type A and B, which is consistent with the SIRR resonant results. Moreover, its overall size is still keeping compact with the size reduction of 56% as the same as discussed before.

Fig. 5.15 Simulated S-parameters of the two optimized dual mode filters
(a) Narrow passband illustration (b) Wide band illustration

5.5 Experimental Results

5.5.1 Filters without Backed-Aperture

In order to verify the simulation results of the optimized dual mode filter, the filter Type A, B and C are fabricated and measured. Unlike the previous experiments, the two ports of the filter are soldered with the SMA connector. The photographs of the circuits are taken after the measurement and removing the SMA connectors, which are demonstrated in Fig. 5.16 (a) and (b). Fig. 5.17 (a) and (b) plot the measured results of these two fabricated filters in the narrow and wide frequency regions, respectively, providing a comparative illustration of their filtering behaviors. The measured results in Fig. 5.17(a) further give us the close view on the electrical performance around the dominant passband.

Fig. 5.16 Photograph of the fabricated filter circuits (a) Type A (b) Type B

As the paired notches in the Type A case are formed on the left-top and right-bottom corners as shown in Fig. 5.16 (a), the two transmission zeros can be observed outside the

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passband at 1.37 and 1.76GHz. Herein these two zeros actually lead to the better improvement of the bandpass behaviors with sharper rejection skirts in the lower and upper stopbands as compared to the other filter Type B. Within the passband of 6.6% at 1.52GHz, the measured magnitudes of S_{11} - and S_{21} -parameters for the two initial prototype filters achieve about 5.5 and 4.5dB, respectively. From the results comparison in Fig. 5.17 (b), it can be seen that the 1st spurious passband appears around 5.30GHz that is 3.49 times of the dominant passband at 1.52GHz in both measured cases. As such, these two filters can really suppress the 1st two parasitic harmonics as existing in the UIRR-based dual-mode filters [53], [54], [66], resulting in significantly broadening the concerned upper stopband in experiment. The frequency shift between the simulated and measured results is mainly caused by the tolerance of the substrate permittivity.

Fig. 5.17 S-parameters of the two fabricated dual-mode filters
(a) Measured S-parameters in the narrow band (b) Simulated and measured $|S_{21}|$ comparison in the wide band

5.5.2 Filters with Backed-Aperture

The dual mode filter Type C is also fabricated and measured in this part and Fig. 5.18

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illustrates the photograph of the circuit configuration: top view layout and the ground plane with cross-shaped aperture. Fig. 5.19 (a) and (b) describe S-parameters of the filter Type C in the narrow passband and wide frequency range respectively. It has the central frequency of 1.52GHz with bandwidth 9.2% in the measured results of Fig. 5.19 (a), and the return loss is higher than 7dB and insertion loss is lower than 2dB in the passband from 1.46-1.60GHz. Zooming out to the results comparison in the wide frequency range in Fig. 5.19 (b), the upper rejection band is verified to be 1.7-5.8GHz in the measured one with insertion loss higher than 10dB. Again, this meander-shaped dual-mode bandpass filter with backed-aperture is experimentally verified the good performance in passband and harmonic suppression with miniaturized circuit size except the frequency shift with similar reason in Type A and B.

Fig. 5.18 Photograph of the fabricated circuit Type C
(a) Top view (b) Bottom view: ground plane with cross-shaped aperture

Fig. 5.19 S-parameters of the SIRR dual-mode bandpass filter Type C
(a) Measured S-parameters in narrow-band view (b) Simulated and measured $|S_{21}|$ comparison in wide-band view

5.6 Conclusion

The ring or patch resonator is a good candidate to design the dual-mode bandpass filter. After examining the characteristic of the two orthogonal modes existing in the ring resonator, we know that the dual-mode filter using one-wavelength ring resonator realization must meet the following conditions: a) The input and output ports should be spatially placed in 90° intervals; b) A discontinuity or so-called perturbation element generating a reflected wave against the incident wave should exist in the ring resonator; c) Symmetrical circuit geometry should be applied. In this work, the novel stepped-impedance ring resonator is firstly proposed for improving the upper rejection band characteristic. The first type of the dual-mode filter using this SIRR and slits perturbation has a wide bandwidth but not so good stopband due to the slit structure. Therefore a class of meander-shaped SIRR is proposed to explore the improved dual-mode filters with miniaturized size and broadened stopband. The numerical results demonstrate that the 1st

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harmonic resonant frequency of this SIRR is raised beyond the three times of its fundamental counterpart. Under the fixed central frequency of the passband at 1.52GHz, the optimized dual-mode bandpass filters with or without backed-aperture adding to the meander-shaped SIRR are able to suppress higher order harmonics and reduce the circuit size effectively. The unexpected tolerances in the substrate parameters and fabrication should be taken into consideration towards further improving the passband behaviors with large return loss and small insertion loss.

Chapter 6

Bandpass Filters Using OLR

6.1 Introduction

Microstrip closed or open-loop resonators (OLRs) with circular or square configuration have been extensively utilized in the design of various microwave components besides its early application in microwave measurement. Particularly, the microstrip open-loop resonator was recently used in exploring a new type of cross-coupled planar bandpass filters [83] response due to its compact size. In order to excite two transmission zeros and properly allocate them on the lower and upper sides of the dominant passband, a multistage filter using open-loop resonators was presented by tapping the input and output feed lines to the asymmetrical position of the first and last open-loop resonators [87]. In

parallel, a microstrip closed-ring or loop resonator sided-coupled to the microstrip feed line was initially investigated in [89], [90] and then applied to minimize the excitation of the 2nd resonant mode in the closed-loop resonator for harmonic-suppressed filter design [91], [92].

In this chapter, a side-coupled microstrip open-loop resonator with half a wavelength ($\lambda/2$) is applied to filter design. After the principle of this simple technique is described in Chapter 2, the two- and four-stage open-loop filters, side-coupled to the microstrip feed lines at two ports, are comparatively studied with varied feed line lengths. Then, the two- and four-stage bandpass filters at 2.52-2.55GHz are optimally designed towards minimizing the 1st spurious passband around 5.10GHz as well as reasonable passband in the first dominant band. At last, the two filter samples are fabricated on the substrate of RT/Duroid 6010 with $\epsilon_r=10.8$ and $h=0.635\text{mm}$ and their performances are measured to provide an experimental proof on the predicted results.

6.2 Two-Stage Open-Loop Resonator Filters

In this section, a two-stage filter, as depicted in Fig. 6.1, is initially constructed by oppositely cascading the two side-coupled microstrip open-loop resonators through the inter-stage capacitive coupling. Using the Agilent ADS simulator, this two-stage filter is numerically simulated under varied parameters such as coupling gap g_1 , g_2 , strip width W and feed line lengths (D). Let's first look at effect of the coupling gap on the filtering

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passband performance.

Fig. 6.1 Layout of two-stage open-loop resonator filter

It is obvious that the gap width is inversely proportional to the coupling degree. The dimensions of the open-loop resonator are selected so that all the filters have the center frequency at 2.4-2.5GHz in this chapter. Fig. 6.2 (a) and (b) illustrate the frequency response of the two-stage filter under different gap widths g_2 , i.e. the coupling degree between the two resonators. The external coupling gap g_1 is fixed to be 0.1mm on this condition. It is seen that the two transmission poles in S_{11} -parameters get separated when the gap g_2 is reduced in Fig. 6.2 (b). It is because the internal mode coupling is getting stronger and is matched with the external coupling in the solid line of $g_2=0.65\text{mm}$. On the other hand in Fig. 6.3 of the fixed the inter-resonator space g_2 , the smallest gap of $g_1=0.1\text{mm}$ in the three cases has a reasonable passband performance for the strong external coupling makes the resonant modes fully coupled to each other.

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Fig. 6.2 Frequency responses of the two-stage filter under different inter-resonator couplings (a) S_{21} -magnitude (b) S_{11} -magnitude

Fig. 6.3 Frequency responses of the two-stage filter under different external couplings (a) S_{21} -magnitude (b) S_{11} -magnitude

After making a good passband performance, the feed line length D must be adjusted to suppress the harmonics occurring at the second multiple of the fundamental frequency f_0 . The numerical simulated S_{21} -parameters of the two-stage filter under varied feed line lengths (D) are obtained. The guided-wavelength at $f_0=2.45\text{GHz}$ is found about 45.5mm, thus deriving $\lambda_{g0}/8 \approx 5.69\text{mm}$. Table 6.1 tabulates the relevant maximum S_{21} -magnitudes,

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$|S_{21}|$, around $2f_0=4.9\text{GHz}$ in the cases of $D=4.0, 5.0, 5.4, 6.0, 6.6, 6.4, 7.0, 8.0$, and 9.0mm , respectively. In the two extreme cases of $D=4.0$ or 9.0mm (much shorter or longer than $\lambda_{g0}/8$), the spurious passband is significantly excited with the maximum $|S_{21}|$ of -4dB or -2dB . As D is adjusted to around $\lambda_{g0}/8$, $|S_{21}|$ is gradually reduced to achieve -51dB at $D=6.0\text{mm}$ and consequently effectively diminishing the unwanted 1st harmonic passband. But, due to the open-end and bend discontinuity effects, the most suitable line length (D) with minimized $|S_{21}|$ is slightly stretched from 5.69 to 6.0mm or from $12.5\%\lambda_{g0}$ to $13.2\%\lambda_{g0}$.

Table 6.1 Simulated two-stage filter S_{21} -magnitude at $2f_0$ versus feed line length D

$D(\text{mm})$	4.0	5.0	5.4	6.0	6.4	7.0	8.0	9.0
D/λ_{g0}	8.8%	11.0%	11.9%	13.2%	14.1%	15.4%	17.6%	19.8%
$ S_{21} @2f_0$	-4dB	-10dB	-15dB	-51dB	-18dB	-7dB	-3dB	-2dB

Fig. 6.4 shows the frequency response of $|S_{21}|$ at $D=6.0, 6.4$ and 9.0mm . We can clearly observe here that $|S_{21}|$ within the 1st harmonic is consistently reduced below than -51dB with a continuous flat curve as D is shortened to 6.0mm . It is found in our simulation and the listed data in Table 6.1 that this harmonic passband becomes increasingly serious when D is further shortened.

Fig. 6.4 Simulated frequency responses: S_{21} -magnitude of a two-stage microstrip open-loop resonator filter with varied line lengths (D)

The above open-loop filters are based on the uniform width resonator that has the same width with the feeding line. It should be noted that thinner microstrip line coupling has more tightened coupling degree than thicker ones. Bearing this knowledge on mind, the filter could be re-configured by reducing the width of the open-loop resonator as shown in Fig. 6.5 (a). The bandpass filter is optimized with varied width W of three cases: $W=0.6$, 0.4 , 0.2mm as Fig. 6.5 (b) illustrates the simulated S-parameters of the respective cases in the passband. With all the center frequencies are tuned to be the same as 2.55GHz , the bandwidth is increased from 5.5% , 7.8% to 11.0% when the width is reduced, thus realizing a much-broadened bandwidth in the solid line. On the other hand, the insertion losses remain around the level of 2dB in the passbands.

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Fig. 6.5 Optimized two-stage bandpass filters
(a) Re-configured filter layout (b) Comparison of the filters with different resonator width
(Optimized dimensions (mm) for $W=0.2$: $L_1=5.0$, $L_2=7.6$, $L_3=3.08$, $g_1=0.1$, $g_2=0.48$)

The gap spacing consideration is similar with the wider strip case due to the allowable tolerance in fabrication. In the case of $W=0.2\text{mm}$, the feed line coupling length D should be adjusted as it has an effect on the external coupling and the second harmonics as well. The guided-wavelength at $f_0=2.55\text{GHz}$ is found about 44.0mm , thus deriving $\lambda_{g0}/8 \approx 5.5\text{mm}$. As such, Table 6.2 tabulates the relevant maximum S_{21} -magnitudes, $|S_{21}|$, around $2f_0=5.10\text{GHz}$ in the cases of $D=3.6, 5.0, 5.4, 6.0, 6.6, 7.0, 8.4$ and 9.0mm ,

respectively. In the two extreme cases of $D=3.6$ or 9.0mm , much shorter or longer than $\lambda_{g0}/8$, the 2nd spurious harmonics is significantly excited with the maximum $|S_{21}|$ of -6dB or -2dB. As D is adjusted at about $\lambda_{g0}/8$, $|S_{21}|$ moves down to -38dB at $D=6.0\text{mm}$, thereby effectively diminishing the unwanted 1st harmonic passband. Again, due to the open-end and bend discontinuity effects, the most suitable line length (D) with minimized $|S_{21}|$ is slightly stretched from 5.5 to 6.0mm or from $12.5\%\lambda_{g0}$ to $13.6\%\lambda_{g0}$, which is consistent with previous case.

Table 6.2 Simulated two-stage filter S_{21} -magnitude at $2f_0$ versus feed line length D

$D(\text{mm})$	3.6	5.0	5.4	6.0	6.6	7.0	8.4	9.0
D/λ_{g0}	8.2%	11.3%	12.3%	13.6%	15.0%	15.9%	19.1%	20.5%
$ S_{21} @2f_0$	-6dB	-12dB	-18dB	-38dB	-15dB	-9dB	-3dB	-2dB

After the data listing on Table 6.2, three of the typical cases in the table are shown in Fig. 6.6, i.e. simulated S_{21} -magnitude with varied line lengths $D=6.0$, 6.6 and 8.4mm. In the best case of $D=6.6\text{mm}$, S-parameters are not as good as the uniform open-loop width case $W=0.6$ which has adequate suppression of the harmonics at the second resonant frequency. This suppression extent might attributes to the width variation and the stepped impedance effect in the feed lines.

Fig. 6.6 Simulated frequency responses of a two-stage microstrip open-loop resonator filter with varied line lengths (D)

6.3 Four-Stage Open-Loop Resonator Filters

Fig. 6.7 Layout of four-stage open-loop resonator filter (Optimized dimensions (mm) for W=0.2: L₁=5.0, L₂=7.6, L₃=3.14, g₁=0.1, g₂=0.27, g₃=0.6)

Lastly, a four-stage open-loop resonator filter is further formed as illustrated in Fig. 6.7 to improve the bandpass behavior. In this aspect, the mixed capacitive and inductive coupling [33], [83] is introduced between the 1st and 2nd resonators as well as between the

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3rd and 4th resonators. It has been discussed in the two-stage case that the width of the open-loop resonator is one of the factors affecting the coupling degree and then on the passband performance. These related results have been plotted in Fig. 6.8, which illustrates the S-parameters of the four-stage filters with different resonator width W covering the dominant passband. It is found the filters have the respective fractional bandwidth of 3.4%, 6.5% and 9.0% in the case of $W=0.6$, 0.4 and 0.2mm , realizing the target of bandwidth broadening.

Fig. 6.8 S-parameters of the four-stage filter with different resonator width

(a) S_{21} -parameters magnitudes (b) S_{11} -parameters magnitudes

In the case of $W=0.2\text{mm}$, the feed line coupling length D is adjusted to illustrate the second harmonics suppression. The guided-wavelength at $f_0=2.52\text{GHz}$ is found about 44.5mm , thus deriving $\lambda_{g0}/8 \approx 5.56\text{mm}$. Table 6.3 gives the relevant maximum S_{21} -magnitudes, $|S_{21}|$, around $2f_0=5.04\text{GHz}$ in the cases of $D=3.6, 4.2, 5.2, 5.6, 6.2, 6.6, 7.6, 8.6\text{mm}$, respectively. In the two extreme cases of $D=3.6$ or 8.6mm , much shorter or longer

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than $\lambda_{g0}/8$, the 2nd resonant frequency is significantly excited with the maximum $|S_{21}|$ of -3dB or -2dB. As D is adjusted at about $\lambda_{g0}/8$, $|S_{21}|$ is reduced to achieve -41dB at D=6.2mm, therefore effectively diminishing the unwanted 1st harmonic passband. Similarly, due to the open-end and bend discontinuity effects, the most suitable line length (D) with minimized $|S_{21}|$ is slightly stretched from 5.56 to 6.2mm or from $12.5\%\lambda_{g0}$ to $13.9\%\lambda_{g0}$, which agrees with previous cases.

Table 6.3 Simulated four-stage filter S_{21} -magnitude at $2f_0$ versus feed line length D

D(mm)	3.6	4.2	5.2	5.6	6.2	6.6	7.6	8.6
D/ λ_{g0}	8.2%	9.4%	11.7%	12.6%	13.9%	14.8%	17.1%	19.3%
$ S_{21} @2f_0$	-3dB	-4dB	-9dB	-15dB	-41dB	-9dB	-3dB	-1dB

Fig. 6.9 Simulated frequency responses of a four-stage microstrip open-loop resonator filter with varied line lengths (D).

Fig. 6.9 depicts the simulated S-parameters of four-stage open-loop resonator filter with varied lengths D. It can be seen that all of the three cases have a reasonable pass band around 2.52GHz, out side of which there are sharp upper and lower rejection bands. In the

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comparative three sets of curves, there are some excitations at the second resonant frequency 5.12GHz. It is significantly suppressed under 41dB when $D=6.2\text{mm}$. It moved up when D is away from 6.2mm and the second passband almost occurs in the case of $D=8.6\text{mm}$. This may give some idea to the design of dual-passband if the coupling parameters are readily chosen.

Fig. 6.10 Optimally designed two-stage filter block
(a) Fabricated photograph (b) Predicted and measured S-parameters

6.4 Experimental Results

To provide an experimental proof, a two-stage filter sample is firstly fabricated and its performance is measured using the Network Analyzer and test fixture. Fig. 6.10 (a) shows

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the photograph of our fabricated circuit sample. Fig. 6.10 (b) depicts the predicted and measured S-parameters of this two-stage filter. Both of them agree well with each other in the whole range from 1.0 to 7.0GHz. The dominant passband bandwidth achieves about 11.5% at the central frequency 2.55GHz with low insertion loss lower than 1.5dB as described in the zoom in box in Fig. 6.10 (b). The dashed and solid lines both indicate that the insertion losses at the 1st harmonic passband are higher than 35dB, which has reached a great extent of suppression. Moreover, the length of the filter is reduced to about 17mm.

Fig. 6.11 Optimally designed four-stage filter block
(a) Fabricated photograph (b) Predicted and measured S-parameters

The simulated and optimized four-stage open-loop resonator filter is also fabricated and

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measured as Fig. 6.11 (a) illustrates its circuit photograph. Fig. 6.11 (b) depicts the optimized and measured S-parameters. First of all, we can see that both of them are well matched with each other with very sharp rejection outside the dominant passband at 2.52GHz. As shown in the zoomed-in graph box together with wide frequency range in Fig. 6.11 (b), the passband width is about 9.1% with the insertion loss of about 3.0dB. From the comparative study on the perfect and copper strip conductors, it is found that the insertion loss is mainly caused by the conductor loss. Meanwhile, the insertion loss in the 1st harmonic is found about 41.0dB and 37dB in simulation and experiment, respectively. These results show us again that the proposed side-coupled technique can really suppress the 1st harmonic passband in an effective way. Furthermore, the overall length of this four-stage filter only occupies 26mm, i.e. $58\%\lambda_{g0}$, thus realizing the significant reduction in size as comparison to the closed-loop filter [91].

Generally, the unloaded Q factor of microstrip line resonators are usually about a few hundreds as compared to a few thousands of waveguide resonator as discussed in [33]. However, the former ones are always preferred in the wireless communications due to their attractive features such as low cost, light weight, easy integration and so on. This issue can not be well addressed in exception that the filter is implemented using superconductive materials at the expense of high cost and large size as reported in [33]. Under this limitation, the insertion loss of high-order filter becomes gradually high as the dominant passband is narrowed and the filter order is increased, which can be observed

from the results illustration in Fig. 6.10 (b) and Fig. 6.11 (b).

6.5 Conclusion

As has been discussed above, there are multiple resonant modes existing at the integrals of the dominant resonant frequency in uniform $\lambda/4$ -, $\lambda/2$ - and λ -type of resonators. In this chapter, a novel class of side-coupled open-loop resonator bandpass filters with compact size and harmonic suppression is proposed, designed and implemented using the characteristics of the current distribution in different lengths of resonators such as the $\lambda/8$ -, $\lambda/4$ -, $\lambda/2$ -types at different resonant frequencies. By setting the open-end feed line length close to $\lambda_{g0}/8$, the 1st spurious resonance at $2f_0$ of the open-loop resonator can be effectively suppressed. After an intuitive description is made on its operating principle, the two- and four-stage filter is numerically characterized versus varied feed line lengths. Then, both of the two- and four-stage filters based on these open-loop resonators are optimally designed. Two groups of predicted and measured results show us that the 1st harmonic passband can be diminished below 35dB with the use of this proposed side-coupled technique at the input and output ports. In addition, the overall length of a four-stage filter is found to be $58\%\lambda_{g0}$, much shorter than the traditional closed-loop filter due to its open-loop half-wavelength resonator feature and the bent-shaped form.

Chapter 7

Ultra-Wideband (UWB) Bandpass Filters

7.1 Introduction

Since the Federal Communications Commission (FCC) in USA authorized the unlicensed use of UWB (3.1~10.6GHz) band for indoor and hand-held systems in 2002 [94], an increasing interest has been recently aroused in both academic and industrial aspects on exploring various UWB components and systems [95]. As the key blocks, a few UWB bandpass filters have been reportedly studied towards building up a UWB passband with 110% fractional bandwidth at 6.85 GHz. In [100], [101], the parallel-coupled line bandpass filter was constructed using three-line tight-coupling structure, but the reported

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bandwidth only achieved 40%-70%. In [105], a microstrip ring filter with the dual stopbands below 3.1GHz and above 10.6GHz was built up with the passband covering the whole UWB band. However, this filter has several drawbacks, such as unexpected passbands below 3.1 GHz, narrow lower and upper stopbands, large size and complexity in configuration.

In the works of [110], [111], a non-uniform transmission line resonator was constituted with three distinct sections with high impedance in the two sides and low impedance in the center. Its first two resonant modes were utilized together with aperture-backed parallel-coupled microstrip line to achieve a 60%~80% passband with four transmission poles. To further achieve a wide passband covering the whole UWB band, this initial multiple-mode resonator can be properly modified so as to excite and allocate its first three resonant-mode frequencies close to the lower end, center and upper-end of the targeted UWB passband. Plus the aperture-backed parallel-coupled microstrip line, hybrid microstrip/CPW structure [115] also has the enhanced coupling degree. It is constructed toward allocating the coupling peak near 6.85 GHz and then two classes of novel compact UWB bandpass filters are developed with five transmission poles. After optimization design is executed via Momentum software on the RT/Duroid 6010 substrate with $\epsilon_r=10.8$ and $h=0.635\text{mm}$, filter samples are fabricated to experimentally verify all the predicted parameters, i.e., insertion/return loss and group delay, in a wide frequency range containing the interested UWB passband.

7.2 Aperture-Backed Microstrip UWB Bandpass Filter

The three-dimensional layout of the proposed microstrip UWB filter is depicted in Fig. 7.1. It is composed of a non-uniform multiple-mode resonator (MMR) in the central part and two identical aperture-backed parallel-coupled microstrip lines in the two sides. The backside aperture on ground plane is formed not only to tighten the coupling degree of coupled lines and also to realize the specified impedance ratio of side to central sections in the MMR [110]. In constructing this MMR, the two side sections are equally selected as

Fig. 7.1 Three-dimensional layout of the aperture-backed microstrip line UWB bandpass filter

one quarter-wavelength at the central frequency of this UWB passband, i.e., 6.85GHz. Meanwhile, the central section is set as one half-wavelength long at 6.85GHz. As such, the 2nd resonant frequency of this MMR can be readily allocated around 6.85GHz. Furthermore, the impedance ratio of the three sections along the MMR is adjusted in order to excite the 1st and 3rd resonant modes of this resonator at the two frequencies slightly higher and lower than 3.1GHz and 10.6GHz, respectively, which are the lower and upper

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ends of the concerned UWB passband. In the following, quantitative investigation is carried out on the resonance behaviors of this MMR resonator and frequency dispersion of the constituted coupled microstrip line with backside aperture.

7.2.1 Proposed Microstrip Multiple-Mode Resonator

Fig. 7.2 (a) depicts the layout of the proposed microstrip multiple-mode resonator, which is composed of a wide strip line in the center and two narrow strip lines with aperture compensations at the two sides. The input admittance of this symmetrical MMR structure with two open ends can be characterized by means of odd-mode and even-mode method.

Fig. 7.2 Proposed multiple-mode resonator
(a) Top view layout (b) Equivalent transmission line topology

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Fig. 7.2 (b) illustrates the equivalent transmission line topology divided by the symmetrical dashed line. In the case of odd-mode, the symmetrical plane is short-circuited with load admittance $Y_L = \infty$, thus the odd-mode input admittance Y_{io} can be expressed in (7.1). When the circuit is treated as even-mode with open-circuited at the dashed plane, then $Y_L = 0$ and the even-mode input admittance Y_{ie} is expressed in (7.2).

$$Y_{io} = -jY_1 \frac{Y_2 - Y_1 \tan \theta_1 \tan \theta_2}{Y_1 \tan \theta_2 + Y_2 \tan \theta_1} \quad (7.1)$$

$$Y_{ie} = jY_1 \frac{Y_1 \tan \theta_1 + Y_2 \tan \theta_2}{Y_1 - Y_2 \tan \theta_1 \tan \theta_2} \quad (7.2)$$

To meet the resonance condition, the input admittance should be equal to zero. Assuming that $\theta_1 = \theta_2$, the odd-mode frequencies (denoted as $f_1, f_3, f_5 \dots$ in this section) have the respective electrical lengths θ_o in (7.3):

$$\theta_o = \arctan \sqrt{K}, n\pi \pm \arctan \sqrt{K} \quad (K = Y_2 / Y_1, n = 1, 2, 3 \dots) \quad (7.3)$$

And the electrical lengths of the even-mode corresponding to the frequencies (denoted as $f_2, f_4, f_6 \dots$ in this section) are:

$$\theta_e = n \frac{\pi}{2} \quad (n = 1, 2, 3 \dots) \quad (7.4)$$

It is observed that the electrical lengths of the first three resonant frequencies f_1, f_2 and f_3 are $\arctan \sqrt{K}$, $\pi/2$ and $\pi - \arctan \sqrt{K}$ respectively, which is consistent with the discussion in Chapter 2. Consequently the physical length of MMR is about λ_g at the 2nd resonant frequency f_2 (λ_g is guided wavelength at the center frequency $f_c = f_2$ i.e. 6.85GHz). The admittance ratio K in the MMR is adjusted so as to excite the 1st and 3rd resonant

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modes of this resonator at the two frequencies slightly higher and lower than 3.1GHz and 10.6GHz, respectively, which are the lower and upper ends of the concerned UWB passband.

Fig. 7.3 illustrates the three sets of frequency responses of a microstrip-line multiple-mode resonator with different lengths ($2l_2$) of central low-impedance section under all the

Fig. 7.3 Frequency responses of three multiple-mode resonators with backside-aperture in sides

other fixed dimensions with $W=2.3\text{mm}$ and $l_1=4.4\text{mm}$. In the case of $2l_2=13.7\text{mm}$, the first resonant frequency is observed to shift to the lower side of the specified UWB passband. In the case of $2l_2=4.7\text{mm}$, the first two resonant frequencies are slightly moved to the upper side while the 3rd frequency is beyond the upper end of the UWB passband.

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The latter case is actually used in [110] to design a wideband filter using the first two resonant modes. By properly selecting $2l_2=7.7\text{mm}$, we can find from Fig. 7.3 that the first three resonant frequencies are now quasi-equally spaced with each other within the specified UWB passband. This MMR will be used later on to design a UWB bandpass filter.

Table 7.1 tabulates the first three resonant frequencies versus the aperture width W with unchanged $l_1=4.4\text{mm}$ and $2l_2=7.7\text{mm}$. In this table, it is observed that the 1st and 3rd resonant frequencies change in a large extent, both approaching to the center 2nd resonant frequency, as the aperture width W increases. Meanwhile, the 2nd resonant frequency has little variation between 6.65GHz and 6.76GHz and reaches its peak at $W=1.70\text{mm}$. The first three frequencies should be equally spaced in the UWB band with the locations above 3.1GHz, near 6.85GHz and below 10.6GHz to obtain a UWB passband covering 3.1 to 10.6GHz. Accordingly, the three frequencies of 4.16, 6.76 and 9.46GHz under $W=1.70\text{mm}$ can be recognized as the best in all the cases listed in Table 7.1.

Table 7.1 Calculated first three resonant frequencies (f_1, f_2, f_3) versus aperture width W for the MMR resonator in Fig. 7.2(a)

W(mm)	0	0.50	1.10	1.70	2.30	2.90
$f_1(\text{GHz})$	3.84	3.89	4.0	4.16	4.25	4.31
$f_2(\text{GHz})$	6.65	6.67	6.72	6.76	6.73	6.65
$f_3(\text{GHz})$	9.60	9.59	9.57	9.46	9.31	9.12

7.2.2 Aperture-Backed PCML Coupling

The parallel-coupled microstrip line (PCML) was presented in the early years in the filter construction with shortened length, large coupling and spurious suppression. The aperture compensation [116], [119] has the ability to weaken the strip-to-ground coupling and then further adjust the coupling degree of the coupling lines.

Fig. 7.4 PCML coupling structure with backed-side aperture (a) Layout (b) Equivalent J-inverter

Fig. 7.4 (a) illustrates the proposed parallel-coupled microstrip line coupling structure with the backed side aperture on the ground and it is equivalent to the J-inverter with two phases at sides in Fig. 7.4 (b). It is obvious that narrowing down the strip width and the strip gap will increase the coupling degree. For some fabrication tolerance reasons, the strip width and gaps are not able to decrease unlimitedly. In Fig. 7.4 (a), the coupled PCML strip width is fixed to be 0.2mm and gap 0.2mm to demonstrate the effects of the aperture on the ground. The length of the PCML strip is about $\lambda_{g0}/4$ with λ_{g0} the one-wavelength at the center frequency $f_c=6.85\text{GHz}$.

Fig. 7.5 indicates the frequency responses of an aperture-backed coupled microstrip line

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with different aperture widths (W). We can see here that S_{21} -magnitude response tends to rise up and a pair of transmission poles appear in the S_{11} -magnitude response as the width (W) is enlarged to 2.0mm and 3.0mm. As discussed in [110], [112], these two poles can

Fig. 7.5 Simulated S-parameters of the aperture-backed PCML coupling structure

be explained by largely raising the normalized J -susceptance with respect to the two feeders' impedance beyond the unity. This characteristic is important in the filter design for the controlling the transmission poles in the passband. Following the previous in [110], [115], it can be imaged that a wide passband covering the whole UWB band be achieved with five transmission poles if the aperture-backed coupled lines are suitably coupled with such a MMR resonator.

7.2.3 Proposed Microstrip UWB Bandpass Filter

In the design, the strip and slot widths in the coupled microstrip lines are set as 0.2mm and 0.1mm in Fig. 7.6 (a) in order to tighten coupling degree and also relax fabrication tolerance. Meanwhile, the three sections of such a MMR are formed with the lengths of

(b)
Fig. 7.6 Proposed UWB bandpass filter
(a) Layout (b) Predicted S_{21} -magnitude responses of different coupling length

4.3, 7.7 and 4.3mm as well as strip widths of 0.2, 0.6 and 0.2mm, respectively, to realize the above-specified allocation of first three resonant frequencies. Fig. 7.6 (b) plots the

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simulated S_{21} -magnitude of a UWB filter over the wide frequency range (1.0 to 13.0GHz) with respect to the three different coupling lengths ($L=0.1, 1.0$ and 4.3mm) under the fixed aperture width (2.3mm) and length (4.4mm). At $L=0.1$ and 1.0mm with weak coupling, the three resonant frequencies can be clearly seen to be almost equally spaced within the UWB passband. As L increases to 4.3mm , the S_{21} -magnitude in the two regions

Fig. 7.7 Simulated results of S-parameters and group delay

among the three peaks moves up so as to realize an almost flat frequency response near the 0-dB line over the desired UWB band (3.1 to 10.6GHz).

Consequently, our optimized UWB bandpass filter then has the dimensions of the layout shown in Fig. 7.6 (a) with the aperture width $W=2.3\text{mm}$ and coupling length $L=4.3\text{mm}$. Fig. 7.7 plots the simulated magnitudes of S-parameters and group delay. In addition to

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the flat frequency response in S_{21} -magnitude lower than 0.5dB, five transmission poles are generated in S_{11} -magnitude in the UWB passband with return loss higher than 10dB. It is known that the first, third and fifth poles are attributed by the first three resonant frequencies of the initial multiple-mode resonator and the second and fourth ones are brought by the implementation of the strengthened coupling structure: aperture-backed parallel coupled microstrip line.

Group delay is a measure of how long it takes a signal to traverse a network, or its transit time. It is a strong function of the length of the network, and usually a weak function of frequency. It is defined by

$$D(\omega) = -\frac{d\Phi(\omega)}{d\omega} \quad (\text{Group delay}) \quad (7.5)$$

where $\Phi(\omega)$ is the phase of the S_{21} -parameter and ω is the angular frequency. Usually, the group delay of a filter is nearly proportional to its order but is inversely proportional to filter bandwidth. Flatness and consistency of the group delay versus frequency is important in microwave filters. As described in Fig. 7.7, the group delay of the designed UWB filter is flat in the UWB passband and the variation is less than 0.20ns.

7.3 Microstrip/CPW Hybrid UWB Bandpass Filters

In this section, another novel MMR-based UWB bandpass filter, as illustrated in Fig.7.8, is proposed and implemented using the hybrid microstrip/CPW structures. Similar with its microstrip counterpart in previous section, a CPW non-uniform or multiple-mode resonator is formed on the ground plane to excite and allocate the first three resonant modes occurring around the lower end, center and higher end of the concerned UWB

Fig. 7.8 Three-dimensional view of the proposed UWB bandpass filter based on hybrid microstrip/CPW structure

passband. Meanwhile, a surface-to-surface [115] or broad-side [107] coupled microstrip/CPW structure is characterized, aiming to allocate its coupling peak with enhanced extent around the UWB’s center or 6.85GHz. This proposed UWB filter can address the two problematic issues existing in the initial UWB filter [108], i.e., insufficiently tight coupling degree between two side-to-side coupled microstrip line and parasitic radiation loss from a wide strip conductor or patch in the central part of the MMR on microstrip line.

7.3.1 Proposed CPW Multiple-Mode Resonator

Let’s start to construct and characterize a CPW multiple-mode resonator (MMR). As shown in Fig. 7.9, the proposed open-ended MMR resonator on CPW is composed of one

Fig. 7.9 Layout of proposed multiple-mode resonator on coplanar waveguide

central CPW with narrow slot width or low impedance and two identical CPW with wide slot width or high impedance at two sides under the fixed strip width. Same as the

Fig. 7.10 Frequency responses of three multiple-mode resonators on CPW

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microstrip non-uniform resonator in section 7.2.1, Fig. 7.2 (b) also depicts the CPW's equivalent transmission line topology with three cascaded sections. To determine the frequencies of the three resonant modes, the input admittance (Y_i) at one of open ends, looking into the MMR, must be zero, $Y_i=0$. Thus, all those resonant frequencies can be solved from a transcendental equation.

Fig. 7.10 illustrates the three groups of frequency responses of the CPW multiple-mode resonator with different lengths ($2l_2$) of central low-impedance section under all the other fixed dimensions with $S_1=1.10\text{mm}$ and $l_1=4.18\text{mm}$. In the case of $2l_2=14.18\text{mm}$, the first resonant frequencies are observed to shift to the lower edge of the specified UWB passband. In the case of $2l_2=4.18\text{mm}$, the first two resonant frequencies are slightly moved to the upper side while the 3rd frequency is beyond the upper end of the UWB passband. By properly selecting $2l_2=9.18\text{mm}$, we can find from Fig. 7.10 that the first three resonant frequencies are now quasi-equally spaced with each other within the specified UWB passband.

Table 7.2 Calculated first three resonant frequencies (f_1, f_2, f_3) versus slot width S_1 for the MMR resonator in Fig. 7.9

$S_1(\text{mm})$	0.18	0.58	0.98	1.10	1.38	1.78
$f_1(\text{GHz})$	3.67	4.01	4.12	4.13	4.15	4.16
$f_2(\text{GHz})$	7.29	7.10	6.84	6.76	6.60	6.34
$f_3(\text{GHz})$	10.91	10.12	9.55	9.40	9.13	8.74

Table 7.2 tabulates the first three resonant frequencies (f_1, f_2, f_3) versus slot width (S_1) under the fixed slot width of 0.18mm in the middle. As shown in Fig. 7.9, the three

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sections in this MMR are selected in such a way that the middle section has about one half guided-wavelength or $l_2 \approx \lambda_{g2}/2 = 9.18\text{mm}$ and the two side sections have about one quarter guided-wavelength or $l_1 \approx \lambda_{g1}/4 = 4.18\text{mm}$ at 6.85GHz (UWB's center), respectively. As S_1 is widened, the first resonant frequency (f_1) quickly increases at the beginning and then becomes saturated around 4.12~4.16GHz. On the other hand, the 2nd and 3rd ones, f_2 and f_3 , seem quasi-linearly decreased with S_1 in the small and large degree of deviation, respectively. To achieve a UWB passband covering 3.1 to 10.6GHz, the first three frequencies target to be equally spaced in the UWB band with the locations above 3.1GHz, near 6.85GHz and below 10.6GHz. According to this criteria, the three frequencies of 4.12, 6.84 and 9.55GHz under $S_1 = 0.98\text{mm}$ can be recognized as the best in all the cases listed in Table 7.2.

7.3.2 Hybrid Microstrip-to-CPW Coupling

Fig. 7.11(a) shows a hybrid microstrip/CPW surface-to-surface coupling structure that was initially studied in [115] to build up a broadband microstrip-to-CPW transition, which has the frequency-dispersive and enhanced coupling degree. In this structure, the upper microstrip conductor is vertically coupled with the central strip conductor of the lower CPW on ground plane via electromagnetic coupling. Its coupling behavior can be characterized in terms of an equivalent unified J-inverter network as illustrated in Fig. 7.11 (b). The J-inverter admittance in fact represents the coupling extent and its maximum peak is properly allocated near 6.85GHz by selecting the coupling length close to $L \approx \lambda_g/4$

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Fig. 7.11 Surface-to-surface microstrip/CPW coupling structure
(a) Layout (b) Equivalent J-inverter network

in Fig. 7.11(a). Fig. 7.12 is the frequency-dependent S-parameters of this coupling structure under different slot widths (S_1). As S_1 is widened, the J-admittance peak is observed to rise up significantly as also studied in [115]. Plus the two quarter-wavelength resonators [112] (phases: θ_{MS} and θ_{CPW} in Fig. 7.11(b)), the two transmission poles can be excited in the two sides of 6.85GHz when $S_1=1.60$ mm. As a result, a five-pole UWB bandpass filter can be expected to be constructed using the above resonator and coupling elements.

Fig. 7.12 Simulated S-parameters of hybrid microstrip-to-CPW coupling structure

7.3.3 Proposed Hybrid UWB Bandpass Filters

The characteristic impedance of the coplanar waveguide is affected by its central conductor and side slot widths: larger slot results in decrement of the conductor-to-ground coupling thus making the characteristic impedance higher than smaller one does. According to this theory another type of multiple-mode resonator on CPW Type B is

Fig. 7.13 Top views of the hybrid UWB bandpass filters two-layer layouts
(a) Type A (b) Type B

constructed besides the first one Type A in section 7.3.1. Fig. 7.8 has already illustrated the three-dimensional schematic of the proposed hybrid microstrip/CPW UWB bandpass filter Type A, in which the MMR resonator on CPW is electromagnetically coupled to the two microstrip feeding lines via surface-to-surface coupling structures. Fig. 7.13 is the top view of two-layer layout of the two UWB bandpass filters with upper strip conductor in dark color, lower CPW plane in light color and slot region in white color. Different from Type A, the low-impedance section in the middle of MMR Type B has wider central CPW conductor. Fig. 7.14 (a) and (b) shows the simulated S-parameters under the two coupled strip lengths, i.e., 0.06 and 3.70mm of the two filters, in relation to the two distinct cases with very weak and optimized coupling degrees, respectively. In Fig. 7.14 (a), the dashed S_{21} -magnitude in $L_1=0.06\text{mm}$ shows that the first three resonant modes of

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the designed MMR resonator occur at 4.13, 6.76 and 9.40GHz, and they are quasi-equally distributed within the UWB band. This is also applied to filter Type B in Fig. 7.14(b) with the three resonant frequencies at 4.33, 7.01 and 9.54GHz. As both of L_1 and L_2 are properly strengthened to 3.7mm, the two portions of S_{21} -magnitude among the three resonant frequencies continue to move up so as to achieve an almost flat frequency response close to the 0dB-line over the UWB passband. Meanwhile, the S_{11} -magnitude in the UWB passband achieves higher than 13.5dB and 18.5dB in the two cases with the five transmission poles.

Fig. 7.14 Frequency responses of the proposed hybrid UWB bandpass filters with different coupling lengths (a) Type A (b) Type B

7.4 Experimental Verification

7.4.1 Aperture-Backed Microstrip UWB Bandpass Filter

After slight adjustment of certain dimensions is made towards achieving the satisfactory

Fig. 7.15 Fabricated aperture-backed microstrip-line UWB bandpass filter
(a) Top view photograph (b) Bottom view photograph
(c) Measured S-parameter and group delay (ns)

insertion loss and return loss, the aperture-backed microstrip UWB bandpass filter is designed and fabricated. The top view and bottom view of the fabrication photographs are shown in Fig. 7.15 (a) and (b) respectively. Fig. 7.15(c) shows the measured magnitudes

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of the two port S-parameters and group delay. These S-parameters prove that our designed UWB filter has high return loss ($<1.0\text{dB}$) and low insertion loss ($>8.0\text{dB}$) in the ultra-wide passband from 2.9 to 10.6GHz. On the other hand, within the realized UWB passband, the measured group delay parameters vary between 0.45ns and 0.70ns with a maximum variation of 0.25ns, thus implying the good linearity of the proposed UWB filter. Also, the overall length of this five-pole UWB filter is found to approximately equal to one full wavelength at 6.85GHz or 16.5mm in Fig. 7.15(a), which is much smaller than that reported in [105].

7.4.2 Hybrid UWB Bandpass Filters

Fig. 7.16 Photographs of the fabricated hybrid UWB bandpass filters
(a) Type A top view (b) Type A bottom view (c) Type B top view (d) Type B bottom view

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Finally, both of the two microstrip/CPW hybrid UWB bandpass filters are fabricated to provide experimental verifications on the above theoretically predicted frequency responses. Fig. 7.16 (a), (c) and 7.16 (b), (d) are the top- and bottom-view photographs of the fabricated filter blocks of Type A and B. The overall lengths of 16.9mm and 17.06mm are found much smaller than those reported in [103], [105]-[107] under the condition of five poles. Fig. 7.17 (a) and (b) depict the measured S-parameters and group delay. Entirely speaking, the measured S-parameters are in good agreement with the predicted ones in Fig. 7.14 (a) and (b) over the wide frequency range except a little high return loss

Fig. 7.17 Measured S-parameters and group delays (a) Type A (b) Type B

of 9.2dB around 8.5GHz of Filter Type A. Filter Type B has an insertion loss as low as 0.6dB and return loss as high as 13dB in the whole passband. The measured group delay ranges between 0.46~0.74ns, with the maximum variation of <0.30ns in Type A and 0.42-0.76ns with maximum variation of <0.32ns in Type B over the UWB passband.

7.5 Conclusion

In this chapter, two sorts of novel ultra-wideband bandpass filters with five poles are proposed and implemented. The first one is configured on aperture-backed microstrip line structures using a single resonator with multiple resonant modes and parallel-coupled microstrip line with enhanced degree. After the principles of the coupling structure and multiple-mode resonator are explained, a compact UWB filter is designed and fabricated. The derived results exhibit its promising UWB bandpass behaviors with wide passband of 3.1 to 10.6GHz, insertion loss of $< 1.0\text{dB}$ and group delay variation of $< 0.25\text{ns}$. Then, the UWB bandpass filters with hybrid microstrip/CPW structure are presented. Two types of multiple-mode resonators on CPW are configured to properly excite and equally space the first three resonant frequencies within the UWB passband. In the meantime, a tightened surface-to-surface microstrip/CPW with relaxed tolerance is constructed to drive these MMRs at its two sides. As such, the attractive UWB passbands with five transmission poles, insertion loss of $< 1.0\text{dB}$, return loss of $> 9.2\text{dB}$ and small group delay variations of $< 0.32\text{ns}$ are realized in theory and confirmed in experiment. In addition, the proposed UWB filters have very small overall lengths of 16.9mm and 17.06mm respectively that are approximately equal to one full wavelength at 6.85GHz.

Chapter 8

Conclusions

Performance-enhanced bandpass filters need to be created to improve the performance of wireless communication systems. Therefore, new effective techniques have to be investigated to miniaturize overall filter size and suppress the spurious harmonics to a certain extent. The stepped-impedance resonator essentially has the ability to control the positions of the multiple higher order resonant frequencies under the proper adjustments of the impedance ratio and the electrical lengths of respective sections. So SIR is suitable in the design of cascaded or dual-mode bandpass filters with small size, ultra-wide stopband or UWB passband. Side-coupled open-loop resonator with open-end feed line is also capable of canceling the excitation of the harmonics as well as reducing size.

Chapter 8 Conclusions

Thus an improved microstrip SIR resonator was proposed by forming the backed-aperture underneath the central narrow strip section in the simulation and experiment. Compared to the traditional UIR, this novel SIR has not only miniaturized its overall size but also completely eliminated the first three spurious resonances. As the capacitive gap has insufficient coupling strength, novel $\lambda/4$ high- and low- impedance transformers, which can be characterized in terms of frequency-dispersive J-admittance or K-impedance inverters, were proposed to provide adjustable strong external coupling. The aperture-backed SIRs in Chapter 3 and impedance transformers have constructed a class of microstrip three-stage bandpass filters with size-miniaturization and ultra-broad stopband in Chapter 4. The bandpass filter Type A with capacitive gap between the feed line and first/last stage resonator has a relatively small bandwidth. Types B and C using the high- or low-impedance transformer have largely enhanced the external coupling degree, thus having much wider passband bandwidth as well as the ultra-wide stopband.

After investigation of the principle of the two orthogonal modes existing in the ring resonator, we know that the dual-mode filter by means of ring resonator realization design on the following conditions: a) The two feed lines should be perpendicularly oriented; b) A discontinuity or so-called perturbation element generating a reflected wave against the incident wave should exist in the ring resonator; c) There is an existence of symmetrical circuit geometry. The novel meander-shaped stepped-impedance ring resonator was proposed to explore improved dual-mode filters with miniaturized size and broadened

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stopband in Chapter 5. It was demonstrated that the 1st harmonic resonant frequency of this SIRR was increased by more than three times of its fundamental counterpart. With different notch positions, two dual-mode filters were then optimally designed and fabricated with or without transmission zeros occurring at two side skirts of the passband. The cross-shaped backed-aperture adding to the meander-shaped SIRR had further effects on spurious harmonic suppression and size reduction.

The multiple resonant modes existing in the uniform λ -type resonators was also applied to $\lambda/4$ - and $\lambda/2$ - or other types of resonators. Using the current distribution in different lengths of resonators at different resonant frequencies, a novel class of side-coupled open-loop resonator bandpass filters with miniature size and harmonic suppression was proposed, designed and implemented in Chapter 6. By setting the open-end feed line length close to $\lambda_{g0}/8$, the 1st spurious resonance at $2f_0$ of the open-loop resonator can be effectively suppressed. The two- and four-stage filters were numerically characterized versus varied feed line lengths and optimally designed after this resonator operating principle description. The simulated and experimental results show us that the 1st harmonic passband can be diminished below 35dB by use of this proposed side-coupled technique at two ports. Furthermore, the overall length of a four-stage filter is much shorter than the traditional closed-loop filter due to its open-loop half-wavelength resonator feature and the bent-shaped geometry.

In Chapter 7, two sorts of novel ultra-wideband (UWB) bandpass filters with five poles

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were proposed and implemented by use of a single multiple-mode resonator and the aperture-backed parallel-coupled microstrip line or hybrid microstrip/CPW structure. The novel multiple-mode resonators were constructed to appropriately excite and equally space the first three resonant frequencies within the UWB passband. Meanwhile, the tightened aperture-backed PCML and surface-to-surface microstrip/CPW couplings with relaxed tolerance were constructed to drive these MMRs at their two sides. Therefore, attractive UWB passbands (3.1-10.6GHz) with five transmission poles, low insertion loss, high return loss and small group delay variations were realized in theory and confirmed in experiment. In addition, all of the proposed UWB filters have very small overall lengths approximately equal to one full wavelength at 6.85GHz. These obtained filters compare very favorably with previous ones.

Overall, the cascaded multi-stage bandpass filters using SIR or OLR, dual-mode bandpass filters using SIRR and UWB bandpass filters using MMR have been designed and fabricated with size-miniaturization in the course of this Ph.D thesis. The enhanced-performances: wide passband, ultra-wide stopband or UWB passband have been achieved and verified by the experimental measurements. At last, there are still some improvements that could be made on the impedance transformer, dual-mode filter and UWB filter:

1. Although the impedance transformer has an excellent coupling degree, it has a disadvantage of large length thus increasing the whole circuit size. It needs to improve these coupling structures with small size using the meander line or folded line approach.

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2. As for the dual-mode filter, one may need to consider further reducing its overall size, widening its upper stopband and raising its power-handling efficiencies for applications in wireless communication.
3. As for the proposed UWB bandpass filters, it needs to be designed to achieve wide low-stop and high-stop bandwidths using the combination of highpass and lowpass filtering approaches.
4. Lastly, all the developed filter blocks may be implemented using advanced fabrication process, such as Low-Temperature Co-fired Ceramic (LTCC), for exploration of low-cost, small-size and light-weight wireless systems.

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