Miniaturized planar RF integrated filters with good harmonic suppression

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Miniaturized Planar RF Integrated Filters with Good Harmonic Suppression

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School of Electrical & Electronic Engineering

A Thesis submitted to Nanyang Technological University
in fulfillment of the requirement for the degree of
Master of Engineering

2005

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STATEMENT OF ORIGINALITY

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

11/Apr/2005

Date

Shi Huamin

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SUMMARY

SUMMARY

Coplanar waveguide (CPW) Filters have been widely considered as one of key building blocks in exploration of RF and microwave integrated circuits. Due to the CPW uniplanar configuration, many circuit elements such as capacitor, inductor and resonator can be easily built up for innovative design of various printed and integrated filter blocks without needing any external packaged or unpackaged lumped elements.

Firstly, the uniform CPW transmission line is overviewed in comparison to its microstrip counterpart in terms of geometrical & electrical features. After the closed-form formulas on CPW are given, the Q-factor of transmission line resonators and parameters of lumped capacitive/inductive coupling elements are described in theory.

Secondly, CPW structures with periodical loading of series-inductive T-shaped slots on the central strip and/or ground planes are investigated, designed and implemented to exhibit the bandgap characteristics of these so-called electromagnetic bandgap (EBG) structures. After numerical analysis is executed using the Momentum simulator, several circuit samples are fabricated to provide an experimental validation on the enhanced bandgap properties of the proposed CPW EBG structures.

Thirdly, various novel half- and quarter-wavelength CPW resonators with modified shapes are theoretically characterized toward simultaneously achieving the size reduction and harmonic-resonance suppression in the filter design. In this aspect, a number of

SUMMARY

simulated S-parameters versus varied dimensions are provided to quantitatively exhibit the drop-off of the fundamental resonant frequency and the expansion of the stopband between the first two resonant frequencies. In addition, the Q-factors of each of these CPW resonators are extracted from those calculated S-parameters and they are tabulated at resonant frequencies to give a direct insight into their loss properties.

Next, several CPW bandpass filters with different configurations of quarter-wavelength CPW resonator and lumped/distributed coupling structures are optimally designed and experimentally implemented. Considering that the coupling degree of traditional lumped CPW coupling elements are not sufficiently high as specified in practical filter design, the quarter-wavelength impedance transformers with increased and decreased characteristic impedances are proposed and constructed as the novel distributed K- and J-inverter coupling structures with enhanced extent. In this aspect, their coupling degrees are extensively characterized in theory to exhibit their attractive coupling enhancement against their lumped-element counterparts. In addition, the concept of stepped-impedance resonator (SIR) is utilized to make up a quarter-wavelength CPW resonator with size reduction and harmonic suppression. With the use of these CPW resonators and coupling structures, several CPW filters with two or four stages are designed, fabricated and measured. Both predicted and measured results exhibit that a relatively wide passband (bandwidth > 20%) is achieved with extremely widened stopband in the upper band.

In final, the conclusions are made on the works involved in this M. Eng. project and some recommendations are provided on the future works relevant to the scope of this project.

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CHAPTER 1: INTRODUCTION

CHAPTER 1: INTRODUCTION

Various planar circuits are widely used for applications in microwave and millimeter integrated circuits due to their small size, light weight and easy integration with other passive/active components [1]. After extensive work was done in the past several decades, microstrip line (MSL) and coplanar waveguide (CPW) have been commonly considered as the most useful transmission lines for constructing these planar circuits. Their fundamentals are briefly described as below for comparison.

1.1 Microstrip line

Microstrip line (MSL) is one of the most popular types of planar transmission lines. It has been extensively used in exploring various hybrid monolithic microwave integrated circuits (HMMICs) for low power microwave signal processing [2].

Its geometrical structure is illustrated in figure 1.1. The thin strip conductor of width W is printed on the top surface of a grounded dielectric substrate with thickness h and relative permittivity ϵ_r .

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Figure 1.1: Microstrip line geometry

This MSL has many distinct features: low profile, size compactness, low fabrication cost, good repeatability in mass production and easy integration with other passive/active circuits. In addition, a systematic set of closed-form design formulas have been developed to explicitly model various MSL-based circuit structures.

As can be seen from the Figure 1.1, MSL line is not a truly planar transmission line as the ground plane is not located on the same surface of signal strip [3]. In order to realize direct connection of the strip conductor and ground plane in shunt, via holes are definitely required in practical designs of MSL structures. In the high frequency region, this requirement leads to increased difficulty in fabrication with high accuracy. Also, the MSL's characteristic impedance and effective dielectric constant are undesirably sensitive to varied substrate thickness due to inhomogeneous medium. More importantly, MSL cannot support the excellent transverse-electro-magnetic (TEM) mode especially at high frequencies, but the dominant mode in this MSL still can be considered as the quasi-TEM mode at low frequencies. Figure 1.2 sketches the E and H field lines of such a quasi-TEM mode in the MSL's cross-section. This quasi-TEM mode has most of field

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lines or energies in the dielectric region underneath the strip conductor in exception that a small fraction of the fields are scattered in the air region. The phase velocity (v_p) of the TEM mode in this inhomogeneous medium differs from its free-space velocity (v_c) [2]. Moreover, they are related with each other via effective dielectric constant (ϵ_{eff}) in such a way that ($v_p^2 = v_c^2 / \epsilon_{eff}$). Intuitively speaking, such a difference, caused by inhomogeneous air-dielectric regions, makes the MSL theoretically impossible to attain the purely TEM mode propagating along the MSL.

Figure 1.2: E and H fields of quasi-TEM mode in microstrip line

1.2 Coplanar Waveguide

The coplanar waveguide (CPW) was first proposed by C. P. Wen in 1969 [4]. A conventional CPW on a dielectric substrate consists of a central strip conductor and infinite ground planes at the two sides on the same surface of this substrate as shown in Figure 1.3.

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Figure 1.3: CPW transmission line geometry

Since most of the energy is distributed in the two slot regions between the central conductor and ground planes, this planar CPW can support a quasi-TEM mode of propagation along the longitudinal direction. It offers several attractive advantages over the above-discussed microstrip line (MSL) [1-5]. Figure 1.4 illustrates the E and H field lines of this even-symmetrical quasi-TEM mode with less frequency dispersion in a wide frequency range, usually named as the CPW mode. In this case, the symmetrical plane at center is the perfectly magnetic wall.

Figure 1.4: E and H fields of even-symmetrical CPW mode

CHAPTER 1: INTRODUCTION

In parallel, the other dominant mode, namely, coupled slot line (CSL) mode can also propagate this CPW with very low cutoff frequency. The E and H field lines of this CSL mode are depicted in Figure 1.5, in which the central plane becomes the perfectly electrical wall due to the odd-symmetrical field distribution.

Figure 1.5: E and H fields of odd-symmetrical CSL mode

In this Master research project, only the symmetrical CPW structures are considered so as to avoid the excitation of the problematic CSL, thus simplifying the design procedure.

In the following, the main advantages of CPW transmission lines are summarized.

- Easy fabrication, easy mounting of shunt/series active and passive devices, and no need for wrap around and via holes.
- Lower radiation loss and less frequency dispersion.
- Not very sensitive on substrate thickness, where the characteristic impedance is basically dominated by the ratio of signal strip width and slot width.
- Very weak cross-talk effects between adjacent CPWs relying on the ground plane, thus realizing easy implementation of high-density microwave integrated circuits.

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- Capable of simultaneously supporting both even- and odd-order modes, thereby receiving a wide application in the balun-based microwave integrated circuits.

So far, various CPW structures have been characterized and employed for applications in directional couplers, mixers, switches, antennas, filters and so on [5].

1.3 Motivation

In the recent years, the demand in exploring miniaturized and high-performance microwave components has been rapidly increased for wireless communications. In this aspect, planar filters work as a key component in the whole RF frond-end module and they have been requested to have several advantageous features such as miniaturized size, harmonic suppression and wide bandwidth.

Based on the better performances of coplanar waveguide (CPW) than that of microstrip line (MSL), this Master's thesis is proposed to focus itself on theoretical modeling, optimal design and practical implementation of a variety of CPW filtering circuit blocks, including the lowpass or bandstop filters using the electromagnetic bandgap structure as well as bandpass filters using the lumped/distributed coupling structures and half-/quarter-wavelength stepped-impedance resonators. In the past, two initial types of simple half-wavelength or $\lambda/2$ CPW filters were developed in [1, 6] and their layouts illustrated in Figure 1.6 (a) and (b). In these two CPW filters, the length of each resonator is selected as one half- wavelength, i.e., $\lambda/2$, at the desired frequency, so as to construct the equivalent shunt or series LC resonators in the dominant passband [7].

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Figure 1.6: Half-wavelength bandpass filters. (a) Capacitive-coupled bandpass filter;
(b) Inductive-coupled bandpass filter

However, this half-wavelength resonator also resonates at its 2nd-order resonant frequency with one full wavelength, thus leading to the quite poor stopband performance [1, 6]. Moreover, the widely-used lumped CPW coupling elements has been illustrated to have the very weak coupling degree. As a result, the developed CPW bandpass filters to date usually suffer from narrow passband and high insertion loss [1, 5-6].

In summary, there are several shortcomings of the above two types of half-wavelength bandpass filters, such as large size, spurious harmonics and narrow dominant passband. To address these problematic issues, a number of improved CPW filters are proposed and developed in this thesis by using the miniaturized quarter-wavelength resonators and distributed CPW coupling structures with enhanced coupling degree.

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1.4 Objective of the Project

In this project, fundamental characteristics of various CPW structures are studied for exploration of advanced CPW filter blocks. At first, the periodic CPW structures with inductive loading are characterized to build up an improved class of electromagnetic bandgap (EBG) structures with broadened stopband. Secondly, a variety of CPW inductive-coupling and capacitive-coupling structures are investigated to demonstrate their coupling characteristics. Thirdly, various CPW resonators with half- and quarter-wavelength are studied to show their multiple resonance behaviors. Next, various compact CPW bandpass filters with quarter-wavelength are optimally designed to show the complete suppression of spurious harmonic passband using the Agilent Advanced Design System (ADS) software. After optimization design, fabrication and measurement are carried out to validate the predicted electrical performances.

1.5 Organization of the Report

The whole Master's thesis is organized as follows:

Chapter 1 briefly describes the advantages and disadvantages of microstrip line (MSL) and coplanar waveguide (CPW) as well as the motivation and objective of this thesis.

CHAPTER 1: INTRODUCTION

Chapter 2 reviews the fundamentals of CPW transmission line, its relevant design formulas and CPW filter designs. Coupling behaviors of Inductive- and capacitive-coupling CPW elements are numerically investigated.

Chapter 3 characterizes the periodic CPW electromagnetic band-gap (EBG) structures and applies them for design of CPW bandstop filters. After the characteristic parameters per unit length are studied, three types of filter blocks are designed, fabricated and tested.

Chapter 4 investigates various quarter- and half-wavelength CPW resonators with varied length and configurations to quantitatively demonstrate their multiple resonance characteristics for bandpass filter design with size reduction and harmonic suppression.

Chapter 5 studies the coupling behaviors of the traditional lumped and the proposed distributed coupling structures and applies them to optimization design of several improved CPW filters with miniaturized size and harmonic suppression. In order to widen the dominant passband, the distributed parallel-coupled and quarter-wavelength impedance-transformer structures are extensively studied toward achieving the coupling enhancement. In final, a number of novel CPW bandpass filters are optimally designed and the predicted results are confirmed by experimental ones of the fabricated circuits.

The last chapter concludes this thesis and discusses the possible future work in the topics relevant to the scope of this thesis.

CHAPTER 2:

LITERATURE REVIEW

2.1 Design parameters for CPW

In CPW, a center strip conductor and two ground planes are separated by a narrow gap at both sides and formed on the same surface of a dielectric substrate as shown in Figure 2.1. The dimensions of the center strip width S , the slot width W , the substrate thickness h and the permittivity of the dielectric substrate $\epsilon_0\epsilon_r$ predominantly determine the effective dielectric constant (ϵ_{eff}), characteristic impedance (Z_o) and the attenuation (α) of this CPW transmission line. [2-5]

Figure 2.1: Geometry of CPW structure

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In the design formulas given as below, the center strip width is set as $2a=S$ and the distance of separation between the two infinite ground planes is indicated as $2b=2W+S$. Thus, the slot width W is equal to $b-a$. The dielectric thickness is designated as h , the relative permittivity of the used substrate is ϵ_r and the thickness of strip conductors is t . In the following analysis, the CPW conductors and the dielectric substrates are assumed to have perfect conductivity and real permittivity, respectively [5]. Hence the structure is considered to be lossless and isotropic.

In this section, many useful design expressions are presented to determine ϵ_{eff} and Z_o of the CPW line with the known dielectric substrate and all the other cross-sectional dimensions. Those formulas were analytically derived using the conformal mapping techniques under the reasonable assumptions that the dominant propagation mode operates as a quasi-TEM mode with only transverse electromagnetic (TEM) fields [5, 8]. In this way, the whole CPW cross section is divided into several homogeneous partial regions. As such, the per-unit-length capacitance in each partial region can be determined separately and the total relevant capacitance can then be calculated as the sum of these derived partial capacitances. In final, the closed-form expressions of Z_o and ϵ_{eff} can be explicitly derived and they are expressed in terms of complete elliptic integrals functions.

$$Z_o = \frac{1}{cC_{air}\sqrt{\epsilon_{eff}}} = \frac{30\pi}{\sqrt{\epsilon_{eff}}} \frac{K(k_o')}{K(k_o)} \quad (2.1)$$

$$\epsilon_{eff} = \frac{C_{CPW}}{C_{air}} = 1 + \frac{\epsilon_r - 1}{2} * \frac{K(k_1)}{K(k_1')} \frac{K(k_0')}{K(k_0)} \quad (2.2)$$

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$$\text{where } k_o = S / (2W + S) = a / b \text{ and } k_1 = \frac{\sinh(\pi a / 2h)}{\sinh(\pi b / 2h)} \quad (2.3)$$

In Appendix A, the detailed equations are given.

Z_o and ϵ_{eff} are basically related to k_o ($=a/b$), thus are dominated by the ratio of the dimensions S and W . In this matter, the size of the CPW circuits can be reduced and enlarged in scale with unchanged parameters if the above ratio is not changed. On the other hand, although the conformal mapping techniques are exactly valid only at zero frequency, they can be very reasonably used for the CPW circuit design even at millimeter wave frequencies if the whole transversal dimension is kept small as compared with the wavelength of the maximum frequency. As usual, the upper frequency limit of the concerned frequency is determined based on the criteria that the effective dielectric constant is not increased so much and the high-order propagation surface-/leaky-wave modes are restricted to the attenuation status [5].

2.2 Q factor of a resonator

In this thesis, the CPW resonators with half- or quarter-wavelength are discussed in Chapter 4. In addition to the resonant frequency, the quality (Q) factor is also a very important parameter of a resonator circuit and it is used to characterize its loss properties if any loss is included in a resonator. The strict definition of Q factor is given as [2, 3, 7]:

$$Q = \omega \text{ (time-average energy stored in system)} / \text{(energy loss per second in system)} \quad (2.4)$$

CHAPTER 2: LITERATURE REVIEW

Planar circuits in opened environment suffer from many aspects of losses, including the substrate loss, conductor loss and radiation loss [5]. The total loss in a transmission line resonator can be in general represented by Q value in the form of $Q = \frac{\omega U}{W}$, where $\omega = 2\pi f_0$, f_0 is the resonant frequency, U is the stored energy and W is the average power loss. This total Q factor can be derived as the sum of the three Q factors, i.e., Q_c , Q_d , Q_r , that are related to the conductor, dielectric and radiation loss, respectively, as pointed put above such that

$$\frac{1}{Q} = \frac{1}{Q_c} + \frac{1}{Q_d} + \frac{1}{Q_r} \quad (2.5)$$

For the conductor loss, if the slot/strip width is selected much larger than the conductor skin thickness, the loss can be effectively minimized [5]. By choosing dielectric material with small loss tangent ($\tan \delta$), the dielectric loss can be also significantly reduced. Furthermore, if the transversal dimension is selected very small as compared to the operating wavelength, the radiation loss in CPW structure can achieve much smaller values than that in the microstrip structure [5]. In this research work, CPW resonator structures are chosen for design of high-performance CPW filters due to high Q factors.

In the case that no external circuitry is introduced, the Q factor is referred as to the unloaded Q factor that reflects the only loss performance of such a resonator. In the real implementation, the resonator is usually driven by the external feed lines via weakly coupling elements [7]. As such, this circuit arrangement causes the lowered overall Q

CHAPTER 2: LITERATURE REVIEW

including the external line effects such that the derive Q factor in this case is named as the loaded Q factor. Figure 2.2 shows the typical frequency response of transmission coefficient (S_{21}) for a two-port resonator circuit with the two coupled ports [7].

Figure 2.2: S_{21} response of a resonant circuit

As illustrated in Figure 2.2, the horizontal line that is dropped off by 3-D from the peak is intersected with the S_{21} curve at the two frequencies, namely, lower and upper cut-off frequencies (f_- and f_+) at the two sides of the central frequency (f_r). These three frequency parameters can be derived in simulation or experiment and they can be used to determine the loaded Q in such a simple formula [7]:

$$Q_L = \frac{f_r}{f_- - f_+} \quad (2.6)$$

If the maximum $|S_{21}|$ at f_r or $|S_{21}(f_r)|$ is obtained, the unloaded Q factor (Q_{unload}) of the considered transmission line can be simply determined as

$$Q_{unload} = \frac{Q_L}{1 - |S_{21}(f_r)|} \quad (2.7)$$

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2.3 CPW Capacitive Coupling Element

Figure 2.3 CPW capacitive coupling element.

(a) layout; (b) π -network; (c) J-model

Lumped CPW capacitive coupling element is constructed by cutting a gap in the inner conductor of the CPW as shown in Figure 2.3(a) and it is also referred as to the end-coupled section in [1, 9-10]. In Figure 2.3 (a), the shaded part represents the strip conductor and ground planes while the remaining part indicates the slot and gap regions.

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Figure 2.3 (b) is the initial equivalent π network with the series susceptance jB_b at center and two identical shunt susceptances jB_a at the two sides. The value of jB_b is determined by the gap spacing between the two CPW open-ends and jB_a is caused by the opened-end effects in the two coupled CPW open-ends. It can be understood that jB_a should be identical at both sides due to the geometrical symmetry along the longitudinal direction.

In order to configure the design of a CPW filter, the circuit model in Figure 2.3 (b) can be further modified as Figure 2.3 (c). This J-inverter network is composed of J-inverter susceptance ($J=B$) and the two equivalent line lengths ($\phi/2$) at both sides. A J-inverter circuit serves as an ideal quarter wavelength impedance transformer or 90° phase shifter [2]. For an instance, the input admittance Y_{in} at port 1 can be expressed in terms of the load admittance Y_{out} by the equation:

$$Y_{in} = J^2 / Y_{out} \quad (2.8)$$

According to the equivalence of the two networks in Figure 2.3 (b) and (c), we can establish the relationship among the parameters of these two networks. The ABCD matrix of the π -network in Figure 2.3 (b) can be written as:

$$\begin{bmatrix} 1 & 0 \\ jB_a & 1 \end{bmatrix} \times \begin{bmatrix} 1 & 1/jB_b \\ 0 & 1 \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ jB_a & 1 \end{bmatrix} \quad (2.9)$$

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After analytical calculation, the above equation can be simplified as

$$\begin{bmatrix} 1 + \frac{B_a}{B_b} & \frac{1}{jB_b} \\ jB_a(\frac{B_a}{B_b} + 2) & 1 + \frac{B_a}{B_b} \end{bmatrix} \quad (2.10)$$

For the network in Figure 2.3 (c), its ABCD matrix can be expressed as below:

$$\begin{bmatrix} \cos \frac{\phi}{2} & jZ_0 \sin \frac{\phi}{2} \\ jY_0 \sin \frac{\phi}{2} & \cos \frac{\phi}{2} \end{bmatrix} \times \begin{bmatrix} 0 & 1/jB \\ -jB & 0 \end{bmatrix} \times \begin{bmatrix} \cos \frac{\phi}{2} & jZ_0 \sin \frac{\phi}{2} \\ jY_0 \sin \frac{\phi}{2} & \cos \frac{\phi}{2} \end{bmatrix} \quad (2.11)$$

Equation (2.11) can be further simplified after matrix multiplication

$$\begin{bmatrix} (BZ_0 + \frac{1}{BZ_0}) \sin \frac{\phi}{2} \cos \frac{\phi}{2} & jBZ_0^2 \sin^2 \frac{\phi}{2} + \frac{1}{jB} \cos^2 \frac{\phi}{2} \\ -jB \cos^2 \frac{\phi}{2} + \frac{1}{jBZ_0^2} \sin^2 \frac{\phi}{2} & (BZ_0 + \frac{1}{BZ_0}) \sin \frac{\phi}{2} \cos \frac{\phi}{2} \end{bmatrix} \quad (2.12)$$

Equations (2.10) and (2.12) must be equal under the equivalence of these two networks, thus deducing the below relationships of the parameters jB_a, jB_b and J, ϕ .

$$\phi = \arctan(B_a Z_o) + \arctan(B_a Z_o + 2B_b Z_o) \quad (2.13)$$

$$JZ_o = BZ_o = |\tan(\phi/2 + \arctan(B_a Z_o))| \quad (2.14)$$

CHAPTER 2: LITERATURE REVIEW

By using the two ϕ and J , any CPW end-coupled structure can be equivalently modeled as a J-inverter susceptance with two certain length transmission lines. In order to design the whole CPW filter circuits, these circuit models can be used to carry out the efficient optimization design based on the equivalent cascaded transmission line networks.

2.4 CPW Inductive Coupling Element

Figure 2.4 CPW Inductive coupling element.

(a) layout; (b) T-network; (c) K-model

CHAPTER 2: LITERATURE REVIEW

Lumped CPW inductive coupling element is made up by electrically direct connection of the center conductor with infinite ground of the CPW [2, 6-7], as shown in Figure 2.4 (a). Figure 2.4 (b) indicates its initial equivalent T-network, consisting of the two identical series reactance jX_a at two sides and a shunt reactance jX_b at center. jX_b is basically determined by the transversal narrow strip width between the conductor and ground. Meanwhile, jX_a is caused by the fringing effects of the CPW short-end.

Similarly, to configure the design of CPW filter, the circuit model in Figure 2.4 (b) need to be further modified as in Figure 2.4 (c) with inductive coupling K-inverter. This K-inverter network is composed of a K-inverter reactance ($K=X$) at center and two equivalent line lengths ($\phi/2$) at both sides. This K-inverter reactance can be used to transform one to the other impedance. As such, the input impedance Z_{in} at port 1 can be related to the output impedance Z_{out} at port 2 by the equation.

$$Z_{in} = K^2 / Z_{out} \quad (2.15)$$

Similarly to the previous case, the two networks in Figure 2.4 (b) and (c) must also be equivalent with each other, thereby being capable of inter-relating their ABCD matrix. For the T-network in Figure 2.4 (b), the ABCD matrix can be written as:

$$\begin{bmatrix} 1 & jX_a \\ 0 & 1 \end{bmatrix} \times \begin{bmatrix} 1 & 0 \\ 1/jX_b & 1 \end{bmatrix} \times \begin{bmatrix} 1 & jX_a \\ 0 & 1 \end{bmatrix} \quad (2.16)$$

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Equation (2.16) can be further simplified as:

$$\begin{bmatrix} 1 + X_a/X_b & jX_a(2 + \frac{X_a}{X_b}) \\ 1/jX_b & 1 + X_a/X_b \end{bmatrix} \quad (2.17)$$

The matrix of the network in Figure 2.4(c) can be initially written as:

$$\begin{bmatrix} \cos \frac{\phi}{2} & jZ_0 \sin \frac{\phi}{2} \\ jY_0 \sin \frac{\phi}{2} & \cos \frac{\phi}{2} \end{bmatrix} \times \begin{bmatrix} 0 & -jX \\ 1/jX & 0 \end{bmatrix} \times \begin{bmatrix} \cos \frac{\phi}{2} & jZ_0 \sin \frac{\phi}{2} \\ jY_0 \sin \frac{\phi}{2} & \cos \frac{\phi}{2} \end{bmatrix} \quad (2.18)$$

and can be further simplified as:

$$\begin{bmatrix} (XY_0 + \frac{1}{XY_0}) \sin \frac{\phi}{2} \cos \frac{\phi}{2} & j \frac{1}{XY_0^2} \sin^2 \frac{\phi}{2} - jX \cos^2 \frac{\phi}{2} \\ \frac{1}{jX} \cos^2 \frac{\phi}{2} + jXY_0^2 \sin^2 \frac{\phi}{2} & (XY_0 + \frac{1}{XY_0}) \sin \frac{\phi}{2} \cos \frac{\phi}{2} \end{bmatrix} \quad (2.19)$$

Under the exact equivalence of the two networks in Figure 2.4(b) and 2.4(c), the below expression can be deduced to express the parameter relationship of these two networks.

$$\phi = \arctan(Y_o X_b) + \arctan(2Y_o X_b + Y_o X_a) \quad (2.20)$$

$$KY_o = XY_o = |\tan(\phi/2 + \arctan(X_a Y_o))| \quad (2.21)$$

CHAPTER 2: LITERATURE REVIEW

Based on the above extensive discussion, lumped CPW open-ended and short-ended structures can be modeled as a J- and K-inverter, respectively, with the two equal transmission line lengths ($\phi/2$). They are utilized in the later chapters to drive the two adjacent CPW resonators or the CPW feed line and CPW resonator at the input and output ports. As such, the $\lambda/2$ and $\lambda/4$ CPW bandpass filters can be constructed with the use of these two types of CPW coupling elements.

CHAPTER 3:

CPW PERIODIC STRUCTURES

Periodic structures with slow-wave and bandstop/bandpass properties were extensively studied several decades ago as discussed in [7] and they have recently been gaining an increase interest in the design of planar circuit devices and components [11-13]. Periodic structures [7] basically refer to the waveguides and transmission lines loaded at periodic intervals with identical obstacles. The loaded objects can be either reactive elements or lumped elements. The main interests on the periodic structures in this chapter are investigated to exhibit the periodic CPW structures with slow-wave and bandstop features. In this chapter, the inductive element is periodically loaded with the uniform CPW line structure in order to build up a stopband filter with broadened bandwidth. After a variety of periodic CPW with inductive loading are characterized, a few CPW periodic structures are optimally designed and practically implemented to validate the attractive bandstop characteristics in experiment.

3.1 Basic CPW inductive loading structure

Figure 3.1 (a) illustrates the layout of the basic inductive loading CPW structure, in which the uniform CPW transmission line is interrupted on the two ground planes by a

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pair of slots of short length L_s . Its equivalent circuit is shown in Figure 3.1 (b) and the attached slot can be simply modeled as a pair of short-ended slotline series stub.

Figure 3.1 Basic inductive-loading structure. (a) layout; (b) equivalent circuit model

Using the equation $Z_{in} = Z_o \frac{Z_L + jZ_o \tan \theta_s}{Z_o + jZ_L \tan \theta_s}$, one can derive the total series impedance

contributed by these paired shorted-end stubs, i.e., $Z_{in} = j Z_o \tan \theta_s$, due to $Z_L = 0 \Omega$.

Herein, θ_s is the electrical length of these stubs and it is readily chosen as the small quantity to derive the positive series reactance or reactive element. As a result, the shorted-end stub can be considered to be series-inductively loaded to the uniform CPW.

This inductive-loaded CPW structure with periodic configuration behaves like a slow-wave structure, thus it can be utilized to make up a class of small-size CPW circuits as discussed in [11-15]. In order to further increase the quantity of such a series inductive element, the longer stub length is always required at low frequencies. However, as the stub length is stretched to be comparable to the operating wavelength, these paired slots produce the increased radiation loss as discussed in [16]. It is because the paired slots can

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be considered as a slot radiator in this case. However, in the CPW circuit design held later on, these paired slot stubs should be designed with enhanced series inductive quantity and reduced radiation loss.

3.2 Unit elements of CPW periodic structures

Unit elements are referred to the unit cells that are directly cascaded to make up the periodic structures. In order to enhance the concerned series inductive quantity with reduced radiation loss in each unit element, the three types of improved CPW unit elements with inductive loading are at first characterized in a comparative manner. Figure 3.2(a) to 3.2(c) describes the geometries of three types of unit elements, namely, Type A, Type B and Type C. Figure 3.2(d) is their equivalent circuit network [11]. These unit elements are formed by etching out a pair of T-shape slots in the central strip conductor and/or ground planes. As such, based on the description in the above section, these paired T-shape slot stubs are basically equivalent to the series inductive reactance at low frequencies. Intuitively speaking, the opposite orientation of electric fields in the two slots of such the three inductive-loading CPW structures can absolutely reduce the unexpected radiation loss as studied in my colleague's Master thesis [17].

As illustrated in Figure 3.2 (d), the paired T-shape slot stubs is equivalently modeled as a π -network with a major series reactance (jX_L) and two identical minor shunt susceptance (jB_L). In this way, the unit elements in Figure 3.2(a) to 3.2(c) with the unit length of T can be considered as the uniform transmission line section with a series inductive obstacle.

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By linking an infinite number of these unit elements together in a cascaded way, various CPW periodic structures can be constructed as implemented in [11-15].

Figure 3.2: Unit elements of CPW periodic structures.

Geometries of (a) Type A, (b) Type B and (c) Type C

(d) Equivalent circuit model

In this chapter, various CPW periodic structures are formed on the dielectric substrate of Duroid 6010 material with $\epsilon_r=10.2$ and $\tan\delta=0.0023$ and thickness=50mil and their electrical performance are characterized using the Momentum involved in the Agilent Advanced Design System (ADS) software. Herein, the 50Ω CPW is constructed with the strip width of $S=50\text{mil}$ and slot width of $W=15\text{mil}$.

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In the numerical analysis, the two-port S-parameters of the three CPW unit elements are directly simulated derived and they are then converted to the ABCD-matrix parameters based on the relationship between S-matrix and ABCD-matrix [2, 7]. As pointed out in [7], the unit element of each CPW periodic structure can be considered as an equivalent uniform transmission line section with the length (T) so that its ABCD matrix can be also expressed as that of the transmission line as below.

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos \theta & jZ_0 \sin \theta \\ jY_0 \sin \theta & \cos \theta \end{bmatrix} \quad (3.1)$$

Under the exact equivalence of the two sets of ABCD matrix, the Bloch-wave or effective characteristic impedance of these CPW periodic structures can be derived in terms of the calculated ABCD parameters in such a simple way,

$$Z_0 = \sqrt{B/C} = \text{Re}(Z_0) + j\text{Im}(Z_0) \quad (3.2)$$

Figure 3.3 shows the derived complex impedance with the real parts in solid lines and the imaginary parts in dotted lines. At the low frequencies, the imaginary parts remain as zero and the real parts rise up slowly. As the frequency increases, the real parts seem to rise up at an exponential speed until the imaginary parts suddenly appear as non-zero quantity. If the frequency further goes up, the imaginary parts become zero and the real parts gradually rise up above the zero. The frequency region with non-zero imaginary part is actually referred as to the electromagnetic bandgap (EBG) [13, 15].

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Figure 3.3 Characteristic impedances of CPW periodic structures

From Figure 3.3, we can see that the low cut-off frequency of this bandgap is gradually moved to the low frequency as the paired T-shape slots are etched in the ground plane, inner strip conductor and then both them. It is because of the gradual enhancement in equivalent series inductive reactance [15]. In the meantime, the upper cut-off frequency is almost unchanged for the 3 types of CPW structure and its value is determined by the unit-cell length or the periodicity of such CPW periodic structures. In simulation, this length is set to the constant of 200mil such that this upper frequency is fixed at 14.0GHz.

Now, the S-parameters of the CPW structure (Type C) with a finite number of unit elements and 50Ω feed lines at sides are characterized to provide a direct view on the bandgap or bandstop properties of interest as studied in [13]. In this aspect, the length (Δ)

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of the paired T-shape slots varies from 60mil to 100mil and 140mil under the fixed unit length of 200mil for comparative study. The relevant results are plotted in Figure 3.4. It can be seen herein that the stopband width is gradually increased with Δ , but the upper cut-off frequency tends to be moved down as D is enlarged beyond 100mil. Therefore, $D=100\text{mil}$ is used in the later simulation.

Figure 3.4: Simulation results for different Δ length with fixed unit number ($N=6$)

In the meantime, the number of unit elements determines the bandstop deepness and sharpness of these CPW periodic structures [13, 15]. To observe this phenomenon, Figure 3.5 is prepared to illustrate the S-parameters under unit-cell number of $N=2, 3, 4$ and 6. We can find here that the CPW structure with $N=6$ can realize the much higher attenuation than the others, achieving the maximum insertion loss of 40dB. Thus, the stopband filters with $N=6$ are further considered and designed in the next section.

CHAPTER 3 CPW PERIODIC STRUCTURE

Figure 3.5: Simulation results for different unit numbers

3.3 CPW Periodic Filter Design

Figure 3.6: Three types of Bandstop filter layouts

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The three bandstop filters are designed using the 3 types of the three above-discussed CPW periodic structures with $N=6$, $\Delta=100\text{mil}$ and $T=200\text{mil}$, in which $\epsilon_r=10.8$ based on our own measurement. Their layouts are shown in Figure 3.7. The two 50Ω CPW feed lines have 50mil strip width and 15mil slot width. In addition, all the T-shaped slots are set as the same width of 10mil .

Figure 3.7: Simulation results of 3 types of bandstop filters

Figure 3.7 depicts the relevant simulation results in comparison. All the three types of CPW filters have the almost same upper cut-off frequency at 13.0GHz due to $T=200\text{mil}$. The bandstop width/depth in type B is much broader than those in type A, illustrating that the series inductance jX_L in strip conductor line is much larger than that in ground planes for the same paired T-shape slots. Moreover, the bandstop width/depth in Type C get slight enhancement over type B because of series cascading of the two series inductances

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in strip conductor and ground planes. To look closely at those curves, we can find that the maximum insertion loss in the stopband achieves -15dB, -38dB and -44dB with widened bandwidth in conjunction with the CPW circuits of Type A, B and C, respectively.

Next, such three bandstop filters with six unit cells or elements are fabricated on the substrate of RT/Duroid 6010 using the simple etching process. Figure 3.8 shows the photos of the three fabricated circuits that are connected with the SMA connectors for experiment using the HP network analyzer. The overall length of these filters is set to 1800mil (feeding line 300mil at each side). The width of the 50Ω feed lines is chosen as 15mil in slot width and 50mil in strip width. The width of the T-shaped slots is kept to be 10mil and they are horizontally placed 10mil apart from the uniform CPW slots.

Figure 3.8: Photos of the fabricated bandstop filters

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Figure 3.9 illustrates the measurement results of these three CPW circuits. As the permittivity is changed from the given 10.2 to our measured 10.8, the experimental results now in Figure 3.9 are well matched with the simulated results in Figure 3.7. But, slight discrepancy can be still observed between the predicted and measured results, in particular, raised return loss in the lowpass band. It is mainly attributed to many unexpected factors such as the CPW-to-coaxial transition, parasitic effect of poor soldering, tolerance in fabrication, non-zero conductor thickness and so on.

Figure 3.9: Measurement results of 3 types of bandstop filters

CHAPTER 4:

CPW RESONATORS

As well known, the resonator element plays a critical role in the design of various RF and microwave bandpass filter blocks. Initially, the uniform CPW resonators with half a wavelength were applied to make up the two basic CPW filter topologies in [1, 6]. Recently, the quarter-wavelength CPW resonators have been presented in [18] to realize the size-reduced and harmonic-suppressed bandpass filters. In this chapter, several modified SIR resonators on CPW [20-23] are constructed using the stepped-impedance resonator (SIR) technique [19] and they are extensively and comparatively studied in theory to demonstrate their attractive performances, such as miniaturized size and broadened upper-stopband.

4.1 $\lambda/4$ SIR resonator

Figure 4.1(a) depicts the original quarter-wavelength ($\lambda/4$) CPW resonator with uniform configuration. This $\lambda/4$ resonator is terminated by open- and short-end at the two sides. The $\lambda/4$ SIR resonator on CPW consists of the two CPW segments with different characteristic impedances as shown in Figure 4.1(b). Its high-impedance segment is arranged in the short-ended size, whereas its low-impedance segment is in the open-end. As such, the frequency spacing or stopband between the 1st two resonant frequencies can

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be widened as discussed in [20]. In the following, such a SIR $\lambda/4$ resonator is investigated via Momentum software to quantitatively demonstrate this enhanced stopband property with respects to the ratios of strip widths and lengths for the two different segments.

Figure 4.1: Layouts of uniform and stepped-impedance resonators.

(a) Uniform CPW resonator; (b) Stepped-impedance resonator.

In the simulation, the two terminals of the CPW resonators are linked with the 50Ω CPW feed lines with the strip width of $S=70\text{mil}$ and slot width of $W=20\text{mil}$ by attaching the inductive (Figure 2.4) and capacitive coupling elements (Figure 2.3) with the strip or spacing width of 20mil as executed in [24]. In the meantime, the overall width in cross section of the CPW resonators is fixed as $2W+S=110\text{mil}$ for comparative study. Also, the total length of the considered uniform CPW SIR resonator has the fixed value of 180mil such that the dominant resonant frequency of about 5 GHz can be achieved. Regardless

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of uniform or SIR configuration, both of the $\lambda/4$ resonators in Figure 4.1 can completely suppress or remove the first harmonic resonance at f_1 , twice the fundamental resonator frequency f_0 as reported in previous works [18, 19 and 24]. Figure 4.2 illustrates the simulated S_{21} -magnitude of this SIR resonator with the fixed strip width of 20mil in the left-side segment versus different strip width (w) of the right-side segment with the constant length of $l=100\text{mil}$. As can be observed in this figure, the 1st resonant frequency (f_0) is largely reduced from 4.9GHz to 3.0GHz and the 1st harmonic resonant frequency (f_1) is slightly raised from 16.0GHz to 16.8GHz, thus increasing the ratio of these two frequencies or upper stopband width to $f_1/f_0=5.45$.

Figure 4.2: Simulation results for various w with $l=100\text{mil}$

Using the technique described in Section 2.2, the loaded and unloaded Q factors at f_0 are extracted from the simulated results in Figure 4.2 and they are tabulated in Table 4.1 with respect to different strip widths (w). The derived unloaded Q -factor values are around

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250-400 and close to those predicted and measured ones reported in [24]. As w is gradually widened, it can be observed here that the loaded and unloaded Q -factors tend to be raised, providing the loss reduction property of the SIR resonator with large impedance ratio.

w	100mil	80mil	60mil	40mil	20mil
Q_L	175	151	130	125	122
$Q_{unloaded}$	381	341	293	280	265

Table 4.1: Loaded and unloaded Q factors of $\lambda/4$ SIR resonator

Figure 4.3: SIR resonator harmonic characteristic

Figure 4.3 shows the two resonant frequencies, i.e., f_0 and f_1 , versus different lengths (L) and widths (w). Herein, the horizontal axis indicates the high-impedance length (L) and

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the vertical axis is the concerned two resonant frequencies. As the width w is increased, the dominant frequency (f_0) is almost linearly shifted lower whereas the harmonic frequency (f_l) becomes a non-linear function of the length (l). It can be clearly observed that the ratio (f_l/f_0) achieves the maximum as l approaches about the half of the total length, i.e., $l=100\text{mil}$, in all the listed cases. In particular, under $l=100\text{mil}$, the ratio (f_l/f_0) gradually increases with the strip width (w) as f_l is raised.

4.2 $\lambda/2$ SIR resonator

In this section, the half-wavelength CPW SIR resonator is considered and characterized. Figure 4.4 shows its layout of such a resonator that is driven at two sides by the feed lines via capacitive coupling gap. For a uniform $\lambda/2$ resonator, the 1st harmonic resonant frequency always is twice of its dominant counterpart. By narrowing the central strip conductor as shown in Figure 4.4, an initial SIR $\lambda/2$ resonator can be constructed to increase the ratio of the above two frequencies so as to achieve the broad stopband [19].

Figure 4.4: Layout of $\lambda/2$ SIR resonator

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In the simulation, the two CPW feed lines have the central strip width of 80mil and the slot width of 20mil with 50 Ω characteristic impedance. The two CPW segments at the two sides of this SIR resonator are set to achieve the low impedance with the strip and slot widths of 90mil and 15mil, respectively. The total resonator length is fixed at 900mil in order to locate the dominant resonant frequency at about 3GHz. In addition, the length of the high impedance segment at center is equal to 500mil that is about half a total resonator length in order to derive the maximum ratio of the 1st two resonant frequencies.

Figure 4.5: Frequency response of $\lambda/2$ SIR resonator

Figure 4.5 shows the simulation frequency response of S_{21} -magnitude with respect to the strip width (X). As X is reduced from 90mil to 10mil, the dominant resonant frequency (f_0) quickly falls down from about 3.0GHz to 2.1GHz and the 1st harmonic resonant frequency (f_1) almost remains same at 6GHz in the first three cases and slightly decreases

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to 5.7GHz for the case of $X=10\text{mil}$. In the last case, the ratio of the two frequencies, i.e., f_1/f_0 , reaches to about 2.8 at maximum. Furthermore, the loaded and unloaded Q factors of this SIR resonator are numerically extracted from the results in Figure 4.5 and they are tabulated in Table 4.2. Compared with the Q -factors of $\lambda/4$ resonator in Table 4.1, we can find that both loaded and unloaded Q -factors of this $\lambda/4$ resonator gain the reasonable enhancement. In particular, the loaded and unloaded Q -factors achieve 236 and 468 in $X=10\text{mil}$, respectively, that corresponds to the maximum ratio of the two impedances for the SIR resonator.

X	90mil	50mil	30mil	10mil
Q_L	159	176	202	236
$Q_{unloaded}$	307	356	409	468

Table 4.2: Loaded and unloaded Q factors of $\lambda/2$ SIR resonator

4.3 T-shaped Resonator

Following the works in Chapter 3, a modified CPW SIR resonator is realized by forming a pair of T-shaped slots on the central strip conductor as illustrated in Figure 4.6. Due to the enhanced series-inductive loading effect, this CPW resonator is expected to widen the concerned the stopband between the first two resonant frequencies and have the reduced radiation loss against the uniform $\lambda/2$ resonator. The detailed dimensions of this resonator circuits are given in Figure 4.6.

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Figure 4.6: Layout of T-shaped resonator

Figure 4.7: Frequency response of T-shaped resonator

The relevant simulated results of frequency response are plotted in Figure 4.7 with respect to different T-shaped slot lengths (T). As T is enlarged from 0mil to 400mil, the dominant frequency (f_0) is reduced from 3.0GHz to 2.5GHz, whereas the harmonic

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frequency (f_l) is almost unchanged at 6.0GHz. As such, the ratio of these two frequencies is increased to 2.5. Also, we can observe from this figure that the transmission zero with the minimum S_{11} -magnitude emerges and just exceeds f_l . This zero is basically caused by the resonance of these attached T-shaped slots. For this reason, the overall length of the T-shaped slots should be kept less than half a wavelength of the slotline transmission line. Table 4.3 tabulates the extracted loaded and unloaded Q factors. As T increases, the relevant Q values are raised from 159 to 223 (loaded Q) and from 312 to 399 (unloaded Q). The studies done in this section imply that this T-shaped resonator have the almost same performances as existed in the traditional SIR resonator discussed in Section 4.2.

T	0mil	200mil	400mil
Q_L	159	182	223
$Q_{unloaded}$	312	347	399

Table 4.3: Loaded and unloaded Q factors of T-shaped resonator

4.4 T-shaped SIR Resonator

In final, a novel CPW resonator is proposed by combining the T-shaped and SIR resonators together and its layout is depicted in Figure 4.8 with the detailed dimensions. As discussed in Section 4.2 and 4.3, one can expect that the eventual series-inductive effect of the central segment can be enhanced so as to further reduce the dominant frequency (f_0) and widen the stopband or ratio of the two resonant frequencies (f_1/f_0).

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Figure 4.8: Layout of T-shaped SIR resonator

Figure 4.9: Frequency response of T-shaped SIR resonator

In analysis, the T-shaped slot length is kept constant, $T=100\text{mil}$, and the central strip width (X) is reduced from 90mil to 50mil at the step of 10mil . Figure 4.9 shows the simulated frequency response of S_{21} -magnitude. It can be seen that the 1st resonant frequency (f_0) is now shifted from 2.4GHz to 2.0GHz while f_l is not changed so much. As

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a consequence, the ratio of the two concerned frequencies is increased to 3, i.e., $f_1/f_0=3$. It implies this proposed CPW $\lambda/2$ resonator can also completely suppress the first harmonic frequency as realized in the uniform $\lambda/4$ CPW resonators.

Table 4.4 indicates the extracted loaded and unloaded Q factors of this T-shaped SIR resonator. The unloaded Q factor gets the large increment and achieves 446 at $X=50\text{mil}$. This T-shaped SIR resonator exhibits the enhanced size-reduction due to the reduced f_0 for the unchanged overall length and reduced transmission loss due to the raised Q factor.

X	90mil	70mil	60mil	50mil
Q_L	224	230	273	252
$Q_{unloaded}$	394	416	492	446

Table 4.4: Loaded and unloaded Q factors of T-shaped SIR resonator

CHAPTER 5

CPW BANDPASS FILTERS

In the above chapters, extensive investigation on various CPW coupling structures and CPW resonator with different configurations has been carried out. In this chapter, the main work is emphasized on the proposal, design and implementation of the miniaturized and harmonic-suppressed CPW bandpass filters using the traditional/improved $\lambda/4$ CPW resonators and the quasi-lumped/distributed CPW coupling structures. Experimental validation is further undertaken by measuring several fabricated CPW filter blocks.

5.1 Lumped-coupling CPW filter

Figure 5.1: Traditional CPW $\lambda/4$ filter. (a) Layout (b) Equivalent circuit model.

CHAPTER 5 CPW BANDPASS FILTERS

Figure 5.1 shows the layout of the traditional 2-stage CPW bandpass filter [20]. The two $\lambda/4$ CPW resonators are linked to the input and output CPW feed lines via lumped K- or inductive-coupling elements, whereas they coupled together via lumped J- or capacitive-coupling element at the center. Figure 5.1(b) is its equivalent circuit model.

As discussed in Chapter 2, the K-coupling element is constructed by directly connecting the strip conductor and ground planes, while J-coupling element is formed by interrupting the two CPW open-ends with air spacing. They can be equivalently utilized to make up the impedance- or K-inverter and admittance- or J-inverter in filter design [1, 6]. Following the work in Chapter 4, one can understand that the CPW $\lambda/4$ uniform resonator have the resonant frequencies that are the odd integers of the 1st dominant resonator frequency f_0 , i.e., $(2n+1)f_0$, where n is the positive integer. As a result, the even-odd resonant modes are completely disappeared [20].

The optimized designs are carried out by adjusting gap widths and shunt inductor widths toward achieving the minimized return loss in the dominant passband. The filters are formed on the substrate of Duroid 6010 with $\epsilon_r=10.2$, $\tan\delta=0.0023$ and $h=50\text{mil}$ and the Agilent ADS Momentum software is used in the whole simulation and design. As shown in figure 5.1(a), the dimension of the two CPW feed lines have the strip width of $S=50\text{mil}$ and slot width of $W=15\text{mil}$ in order to achieve the 50Ω characteristic impedance. To set up the center operating frequency around 5.4GHz, the $\lambda/4$ resonator length is chosen as 200mil. To enlarge the inductive coupling effect at both ends, the relevant strip width is set up to the minimum value of 5mil. In addition, the feed line length is set to 300mil.

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Figure 5.2(a) and (b) show the predicted results in the wide and narrow frequency ranges. Initially, we can observe that 1st harmonic passband appears at the three times of its dominant passband. As such, the lowest harmonic in the $\lambda/2$ CPW resonator is completely removed in the considered $\lambda/4$ CPW resonator. In the passband, the bandwidth is found very narrow and achieves about 2.8%. Meanwhile, the S21-magnitude in the passband is lower than -4dB and it is caused by the limited Q-factor and narrow passband.

(a) Wide frequency range

(b) Narrow frequency range

Figure 5.2: Frequency response of traditional CPW $\lambda/4$ filter.

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In order to reduce this insertion loss and also achieve the passband performance, the inductive coupling degree at the first and last stage must become much larger than the capacitive coupling degree in the center. But, there may be some unrealistic issues as listed below to further narrow the strip width in the lumped inductive coupling element:

1. Allowable fabrication tolerance in etching process require the minimum size not less than 4mil for the strip and slot widths.
2. If the strip width is too narrow, the thickness (0.7mil) of the strip conductor starts to bring out some parasitic effects in addition to the increased conductor loss.

5.2 Overlapped-coupling CPW filter

To solve the above coupling problem, an overlapped-coupling structure was proposed in [26] and used in [25] for CPW filter design. In this section, such an enhanced coupling structure is utilized to design a 2-stage CPW $\lambda/4$ filter with the layout as depicted in Figure 5.3, in which the length L_c , basically control the concerned coupling degree.

Figure 5.3: Layout of overlapped-coupling CPW $\lambda/4$ filter

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In the design, the spacing or strip width between the two overlapped slots is set to 5mil. The strip and slot widths at the feed lines are still kept as 50mil and 15mil as before. As such, those dimensions in the CPW $\lambda/4$ resonator become 90mil in strip width and 15mil in slot width. In addition, the resonator length is set to 200mil in order to achieve the same center frequency of the dominant passband as that in the $\lambda/2$ resonator. After optimization design, the overlapped coupled length at two sides and the gap spacing at center, as shown in Figure 5.3, are found 35mil and 82mil, respectively.

The simulated S-parameters are illustrated in Figure 5.4(a) and 5.4(b) in the wide and narrow frequency ranges, respectively. Again, the 1st harmonic passband in the $\lambda/4$ resonator is completely removed. In the dominant passband, the S_{21} response is improved with reduced insertion loss of about 0.9dB and widened bandwidth of about 10.3% due to the enhanced inductive coupling of overlapped arrangement described above. In addition, we can find that the overall size is largely reduced by the overlapped structure.

(a) Wide frequency range

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(b) Narrow frequency range

Figure 5.4: Frequency response of overlapped-coupling CPW $\lambda/4$ filter.

5.3 Modified overlapped-coupling CPW filter

Figure 5.5: Layout of improved CPW $\lambda/4$ filter.

(a) Layout; (b) Field distribution

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Generally speaking, the capacitive coupling at center should be weak and the inductive coupling should be strong enough to achieve the better passband performance. In Figure 5.3, the large gap spacing of 82mil is set up to achieve the designated weak coupling degree. So, it may cause the enlargement in the overall length of the filter. To further reduce the filter's size, the transversal strip conductor with narrow width is inserted into the coupled gap and it is further directly or electrically connected with the two ground planes at two sides. As such, a modified CPW filter with miniaturized size can be made up and its layout is illustrated in Figure 5.5(a).

As discussed in [27], the coupling degree at center can be largely reduced as the grounded strip is introduced. Figure 5.5(b) shows the electrical or E field distribution over the slot region in the CPW resonators and feed lines. Around the two opened-end regions, the E field is distributed in the transversal direction, implying that the whole slotline length in the $\lambda/4$ CPW resonator is stretched in relevance to the strip width. It can be imaged that the longitudinal length of these resonators should be reduced if the same central frequency of the dominant passband is considered.

In this modified design, the gap width at the two opened-ends is selected as 15mil that is equal to the slot width of CPW resonators. The overlapped coupling length at the input and output sides are kept as 35mil to achieve the unchanged inductive coupling degree. The optimization is carried out toward minimizing the return loss within the dominant passband with the center frequency of 5.2GHz by properly adjusting the length of CPW resonators and the strip width at center.

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(a) Narrow frequency range

(b) Wide frequency range

Figure 5.6: Frequency response of improved CPW $\lambda/4$ filter

As a result, the strip width (d) and resonator length are determined as 10mil and 185mil, respectively. Figure 5.6(a) and 5.6(b) depict the relevant simulated results in the narrow and wide frequency ranges. The passband still has the unchanged bandwidth of 10.2% as that realized in the filter studied in the above section. Furthermore, the insertion loss in

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the dominant passband is slightly reduced from 0.90dB to 0.75dB and that in the harmonic passband is increased higher than 10dB at the cost of increased radiation loss.

5.4 Impedance transformer coupling structures

In order to widen the dominant passband under the allowed fabrication tolerance, the inductive coupling of the above-discussed coupling structures are found not sufficiently large. To overcome this problem, a simple $\lambda/4$ CPW impedance transformer with low characteristic impedance is proposed as a novel class of structures with enhanced inductive or capacitive coupling degree for design of broadband CPW filters [22-23].

5.4.1 $\lambda/4$ impedance transformer (K-coupling)

To carry out the comparative study, the layout of the $\lambda/4$ impedance transformer with low impedance is drawn together with that of the traditional lumped K-coupling structure as illustrated in Figure 5.7(a) and 5.7(b). Figure 5.7(c) shows their unified equivalent K-inverter network with a K-inverter reactance (K) and two transmission line lengths ($\phi/2$) length at both sides. Herein, the K-reactance actually represents the concerned inductive coupling degree of the structures in Figure 5.7(a) and 5.7(b). For the $\lambda/4$ transformer with the length θ and impedance Z_{0l} in Figure 5.7(b), its relevant ABCD matrix can be written:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos \theta & jZ_{0l} \sin \theta \\ jY_{0l} \sin \theta & \cos \theta \end{bmatrix} \quad (5.1)$$

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Figure 5.7: Layouts and circuit models of two K-coupling structures

In the meantime, the ABCD matrix of K-inverter network in Figure 5.7(c) can be derived by cascading those of transmission lines and K-reactance such that

$$\begin{bmatrix} \cos \frac{\phi}{2} & jZ_0 \sin \frac{\phi}{2} \\ jY_0 \sin \frac{\phi}{2} & \cos \frac{\phi}{2} \end{bmatrix} \times \begin{bmatrix} 0 & -jK \\ 1/jK & 0 \end{bmatrix} \times \begin{bmatrix} \cos \frac{\phi}{2} & jZ_0 \sin \frac{\phi}{2} \\ jY_0 \sin \frac{\phi}{2} & \cos \frac{\phi}{2} \end{bmatrix} \quad (5.2)$$

After matrix multiplication, Equation (5.2) can be deduced as a single ABCD matrix.

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$$\begin{bmatrix} \left(\frac{K}{Z_0} + \frac{Z_0}{K}\right) \sin \frac{\phi}{2} \cos \frac{\phi}{2} & j \frac{Z_0^2}{K} \sin^2 \frac{\phi}{2} - jK \cos^2 \frac{\phi}{2} \\ \frac{1}{jK} \cos^2 \frac{\phi}{2} + j \frac{K}{Z_0^2} \sin^2 \frac{\phi}{2} & \left(\frac{K}{Z_0} + \frac{Z_0}{K}\right) \sin \frac{\phi}{2} \cos \frac{\phi}{2} \end{bmatrix} \quad (5.3)$$

By defining the normalized K-reactance $\bar{K} = K / Z_0$ where Z_0 is the characteristic impedance of the transmission lines at two sides in Figure 5.7(c), the above matrix can be further simplified as:

$$\begin{bmatrix} \left(\bar{K} + \frac{1}{\bar{K}}\right) \frac{\sin \phi}{2} & \frac{jZ_0}{2} \left[\left(\frac{1}{\bar{K}} - \bar{K}\right) - \cos \phi \left(\frac{1}{\bar{K}} + \bar{K}\right) \right] \\ \frac{j}{2Z_0} \left[\left(\frac{1}{\bar{K}} - \bar{K}\right) - \cos \phi \left(\frac{1}{\bar{K}} + \bar{K}\right) \right] & \left(\bar{K} + \frac{1}{\bar{K}}\right) \frac{\sin \phi}{2} \end{bmatrix} \quad (5.4)$$

Under the equivalence of the two networks in Figure 5.7(b) and 5.7(c), the two resultant ABCD matrices in Equations (5.1) and (5.4) must be the exactly same. After the effort on analytical work is done, the below two equations can be derived to relate the parameters in these two networks.

$$\tan \phi = -\frac{2}{Z_{0l} + Y_{0l}} \times \frac{1}{\tan \theta} \quad (5.5)$$

$$\frac{1}{\bar{K}} - \bar{K} = (\bar{Z}_{0l} - \bar{Y}_{0l}) \sin \theta \quad \text{or} \quad \bar{K} = -\frac{\sin \theta}{2} \left[(\bar{Z}_{0l} - \bar{Y}_{0l}) + \frac{\bar{Z}_{0l} + \bar{Y}_{0l}}{\cos \phi} \right] \quad (5.6)$$

In particular, when $\theta=90^\circ$, one can obtain $\bar{K} = -\bar{Z}_{0l}$ since $\tan \theta$ is infinity. It implies that K-reactance is exactly equal to the impedance of the transformer with the $\lambda/4$ length.

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Under the consideration of 50Ω CPW feed lines, the normalized K-inverter impedances of the inductive coupling structures in Figure 5.7(a) and 5.7(b) are derived using the Equations (2.21) and (5.6) in order to provide the comparative investigation on their coupling properties. Figure 5.8 shows the derived normalized-K, i.e., $\bar{K} = K/Z_{0H}$, with respect to the frequency under varied gap spacing (D) and slot width (S_L) in conjunction with the above two coupling structures.

Figure 5.8: Normalized K-impedance vs frequency

As can be observed, for the proposed transformer structure, the normalized-K value gradually decreases from the unity value (maximum coupling) around the resonant frequencies as the slot width S_L is reduced from 20mil (uniform transmission line case). This reduced K-value indicates the gradual reduction on the concerned inductive coupling degree and it is caused by the lowered characteristic impedance (Z_{0L}) of the $\lambda/4$

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transformer. However, for the traditional lumped structure, the normalized K increases as a linear function of the frequency and this coupling degree slightly increases as the gap width Δ decreases. Looking at the two different sets of curves in Figure 5.8, the proposed impedance transformer based structure can easily achieve much higher coupling degree than its lumped counterpart.

5.4.2 $\lambda/4$ impedance transformer (J-coupling)

In practical design of CPW filter, the lumped gap or end-coupled capacitive coupling element discussed in Chapter 2 also can not achieves the sufficiently high degree. For this reason, the $\lambda/4$ impedance transformer with high characteristic impedance is proposed and used as an enhanced capacitive coupling element. Its relevant layout is displayed in Figure 5.9(b) together with that of the lumped gap structure in Figure 5.9(a). Figure 5.9(c) indicates the equivalent J-inverter network of the constructed transformer based coupling structure, consisting of a J-inverter susceptance (J) at center and the two equal transmission line lengths ($\phi/2$) of high impedance (Z_{oH}).

Similarly to the discussion in the above section, the ABCD matrix of the transformer structure in Figure 5.9(b) can be simply expressed in terms of its impedance (Z_{oH}) and electrical phase (θ).

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos \theta & jZ_{oH} \sin \theta \\ jY_{oH} \sin \theta & \cos \theta \end{bmatrix} \quad (5.7)$$

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(a) traditional lumped J coupling

(b) Impedance Transformer

(c) Equivalent network

Figure 5.9: Layouts and circuit models of two J-coupling structures

In the meantime, the overall ABCD matrix of cascaded topology in Figure 5.9(c) can be initially expressed as the multiplication of the three ABCD sub-matrices,

$$\begin{bmatrix} \cos \frac{\phi}{2} & jZ_0 \sin \frac{\phi}{2} \\ jY_0 \sin \frac{\phi}{2} & \cos \frac{\phi}{2} \end{bmatrix} \times \begin{bmatrix} 0 & 1/jB \\ -jB & 0 \end{bmatrix} \times \begin{bmatrix} \cos \frac{\phi}{2} & jZ_0 \sin \frac{\phi}{2} \\ jY_0 \sin \frac{\phi}{2} & \cos \frac{\phi}{2} \end{bmatrix} \quad (5.8)$$

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and then becomes such an eventual ABCD matrix,

$$\begin{bmatrix} (BZ_0 + \frac{1}{BZ_0}) \sin \frac{\phi}{2} \cos \frac{\phi}{2} & jBZ_0^2 \sin^2 \frac{\phi}{2} + \frac{1}{jB} \cos^2 \frac{\phi}{2} \\ -jB \cos^2 \frac{\phi}{2} + \frac{1}{jBZ_0^2} \sin^2 \frac{\phi}{2} & (BZ_0 + \frac{1}{BZ_0}) \sin \frac{\phi}{2} \cos \frac{\phi}{2} \end{bmatrix} \quad (5.9)$$

By defining the normalized J-inverter susceptance ($\bar{B} = BZ_0$), Equation (5.9) can be further simplified as

$$\begin{bmatrix} (\bar{B} + \frac{1}{\bar{B}}) \frac{\sin \phi}{2} & \frac{jZ_0}{2} \left[(\bar{B} - \frac{1}{\bar{B}}) - \cos \phi (\frac{1}{\bar{B}} + \bar{B}) \right] \\ \frac{j}{2Z_0} \left[(\frac{1}{\bar{B}} - \bar{B}) - \cos \phi (\frac{1}{\bar{B}} + \bar{B}) \right] & (\bar{B} + \frac{1}{\bar{B}}) \frac{\sin \phi}{2} \end{bmatrix} \quad (5.10)$$

Under the equivalence of the two circuit networks in Figure 5.9(b) and 5.9(c), the ABCD matrix in Equation (5.7) must be exactly equal to that in Equation (5.10). As a result, the relationship between the parameters of these two networks can be analytically derived as,

$$\tan \phi = -\frac{2}{Z_{0h} + Y_{0h}} \times \frac{1}{\tan \theta} \quad (5.11)$$

$$\bar{B} - \frac{1}{\bar{B}} = (\bar{Z}_{0h} - \bar{Y}_{0h}) \sin \theta \quad \text{or} \quad \bar{B} = \frac{\sin \theta}{2} \left[(\bar{Z}_{0h} - \bar{Y}_{0h}) - \frac{\bar{Z}_{0h} + \bar{Y}_{0h}}{\cos \phi} \right] \quad (5.12)$$

Looking at the above equations, we can find that at the frequency where the phase is exactly equal to $\theta = 90^\circ$ of this transformer, the normalized J-susceptance is strictly equal to the characteristic admittance normalized to that at the feed line.

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Figure 5.10: Normalized J-admittance vs frequency

Figure 5.10 depicts the magnitudes of the normalized J-inverter susceptance of the two coupling structures in Figure 5.9(a) and 5.9(b). Let's look at the transformed related results at first. In this case, the normalized J-magnitude gradually falls down around the center frequency and it tends to decrease from the unity value (maximum coupling case) to the small value as the slot width (S_H) is enlarged or its corresponding impedance is increased in the transformer section. On the other hand, for the traditional lumped structure, the J-value shows a linear increase with frequency and the related increment slope is slightly raised as the gap spacing (Δ) is increased. However, the coupling degree of the lumped gap is always much smaller than that of the transformer based one.

In the rest sections of this chapter, these two proposed transformer based coupling structures are utilized to make up the two improved classes of the CPW bandpass filters

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with relatively wide dominant passband and good harmonic suppression. The whole design procedure is carried out using the Momentum software by properly adjusting several dimensions included in the filter structures.

5.5 K-coupling filter with uniform/SIR resonator

Figure 5.11(a) indicates the layout of a 2-stage CPW bandpass filter. The low-impedance transformer sections, linked with the 50Ω feed lines at the two ports, act as the inductive coupling elements with enhanced degree. The two $\lambda/4$ CPW resonators with narrow signal strip width is capacitive coupled at center via weakly end-coupled gap. Figure 5.11(b) depicts its relevant equivalent circuit network, in which the transformer sections are expressed into the K-inverter network as discussed above.

Figure 5.11: Layout of 2-stage CPW filter with uniform resonators

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For simplification in design, the overall transversal width, including the central strip and two slots, are selected as the constant of 110mil. The CPW feed lines have the strip width of 70mil and the slot width of 20mil to achieve the 50Ω impedance. As discussed above, the transformer based coupling structure usually has the extremely large coupling degree in opposite to the extremely coupling in the traditional lumped structure. As such, to achieve the relatively high coupling degree, the strip width needs to be widened and/or the slot width needs to be narrowed to a certain degree, as illustrated in Figure 5.11(a), toward properly reducing the relevant characteristic impedance.

Meanwhile, in order not to request the extremely low impedance with very narrow slot width in the transformer section, the slot width of the two $\lambda/4$ CPW resonators is largely increased to achieve the high characteristic impedance. As such, the high impedance ratio of the resonator and transformer can be realized to reduce the normalized K-inverter reactance, thus reducing the coupling degree of this transformer as requested in the design of CPW filter with low insertion loss in the relative wide passband. For this reason, the strip width of the two resonators is set to 20mil width with the slot width of 45mil. In this case, the impedance of the resonator section is found about 90Ω .

At last, the slot spacing (s) of the coupled gap at center between the two $\lambda/4$ CPW resonators are adjusted in order to achieve the minimized return loss within the dominant passband. As indicated in Figure 5.11(a), this spacing (s) is determined as 4mil. As all the dimension of this 2-stage CPW filter is determined, the relevant circuit sample is then fabricated to provide an experimental validation on the predicted electrical performances.

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Figure 5.12: Frequency response of 2-stage uniform resonator CPW filter

Figure 5.12 shows the simulated and measured results in the range of 2.0 to 16.0GHz. Both them are found in very reasonable agreement with each other, showing that the bandwidth of this 2-stage filter is significantly increased to about 28% as compared to 10.4% of the filter discussed above. Again, the 1st harmonic passband in the $\lambda/2$ resonator is observably suppressed with the return loss higher than 10dB in measurement.

In order to further achieve better harmonic rejection, a 4-stage CPW bandpass filter is designed as well and its layout is illustrated in Figure 5.13(a). In this aspect, the two additional $\lambda/4$ CPW resonators that are coupled together via lumped inductive coupling element at center are inserted into the gap region of the two initial CPW resonators. After proper and simultaneous adjustment of the widths of coupled gap spacing and strip conductor, the better filtering performances have been achieved with the wide and deep

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stopband between the dominant and lowest harmonic passband. Figure 5.13(b) shows the optimized and experimental results. The return loss S_{21} in the dominant passband exhibits the existence of the four transmission poles and its magnitude is higher than 15dB within the whole passband as can be observed in both simulated and measured curves. In addition, the dominant passband width of this 4-stage filter is about 28.3% and the insertion loss reaches as high as 40dB in the stopband region.

Figure 5.13: 4-stage K-coupling CPW filter with uniform resonators.

(a) Layout; (b) Frequency Response.

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In order to further widen the stopband with high rejection, the $\lambda/4$ stepped-impedance resonator (SIR) is constructed and further applied to design the 2- and 4-stage CPW filters. Figure 5.14 shows the layout of the 2-stage filter, in which the uniform $\lambda/4$ resonator in 5.11(a) is replaced by the SIR resonator with the two separated CPW sections with different strip/slot widths or impedances.

Figure 5.14: Layout of the K-coupling CPW filter with SIR resonators

Following the design procedure above, the high-impedance strip width in the SIR resonator is fixed to the small constant of 20mil, whereas its low-impedance counterpart is selected as 70mil. Furthermore, the lengths of low- and high-impedance sections are readily equalized as discussed in [19] and they are equal to half an overall resonator length of 300mil, in which the center frequency of the dominant passband is close to 3GHz. Similarly, by properly adjusting the coupled gap spacing and width, the optimized filtering performances with minimized return loss in the passband can be realized. Figure 5.15(a) and 5.15(b) illustrate the predicted S-parameters in the wide and narrow frequency ranges, respectively. At first, the concerned bandwidth is seen to be further increased beyond 35%. Due to the construction of the $\lambda/4$ resonator, the lowest harmonic passband is now raised to the higher value that is about 3.7 times of the dominant one,

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thereby confirming that the SIR resonator can really improve the stopband properties with widened bandwidth and increased return loss.

(a) Wide frequency range

(b) Narrow frequency range

Figure 5.15: Frequency response of 2-stage K-coupling filter with SIR resonators

To further improve the stopband performance, a 4-stage CPW filter is designed as well and its layout is shown in Figure 5.16 (a). In comparison to the filter in Figure 5.13(a), the only significant modification is to replace the four uniform resonators by the four SIR

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resonators. The optimization design is still undertaken adjusting the dimensions of each coupling structure used in this filter toward minimizing the return loss in the passband.

(a) Layout

(b) Frequency response in wide range

(c) Frequency response in narrow range

Figure 5.16: 4-stage K-coupling filter with SIR resonators

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Figure 5.16(b) shows the simulated S-parameters of this 4-stage filter. We can find that the stopband performance is further improved with the insertion loss as high as 60dB and the fractional bandwidth of about 100%, in which the lowest harmonic frequency appears at 11GHz with reference to 3.0GHz of the central passband frequency. In total, the 4-stage filter has higher and sharper stopband performance than its 2-stage counterpart.

5.6 K-coupling filter with modified resonator

In this section, a 2-stage CPW filter with modified SIR resonator is considered and designed, aiming at further increase the bandwidth of the dominant passband. Its layout is shown in Figure 5.17. As discussed above, the transformer based structure has the inherently high inductive coupling degree as large as the unity normalized J-inverter reactance. Meanwhile, the capacitive coupling degree of the end-coupled gap is primarily dependent on the gap spacing. In order to avoid the parasitic effects of conductor thickness and also fabrication tolerance, the minimum gap spacing must be not as small as 2mil. For this reason, the two coupled open-ends are longitudinally narrowed in length and transversely stretched in width in order to make up a tightly parallel-distributed coupling structure. In the following design, the spacing and the length of this parallel-coupled element is set to the constant of 10mil. After optimization design, a circuit sample is fabricated for experimental verification. Figure 5.18 illustrates the predicted and measured results in the wide and narrow frequency ranges. Both results are matched with each other quite well that the bandwidth of dominant passband is further widened to about 46% and the return loss within the passband exceeds 20dB .

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Figure 5.17: Layout of modified K-coupling CPW filter

(a) Frequency response in narrow range

(b) Frequency response in wide range

Figure 5.18: Frequency response for K-coupling filter with modified resonators

5.7 J-coupling filter with SIR resonator

As studied in Section 5.4.2, the $\lambda/4$ CPW transformer with high impedance can achieve the much high capacitive coupling degree. So, this capacitive element is utilized at the input and output ports to construct an alternative class of broadband CPW filter. Figure 5.19 shows the geometrical layout of the proposed CPW bandpass filter, in which the high-impedance $\lambda/4$ transformer with narrow strip width is put at two sides and lumped inductive element with weak coupling degree is utilized to link the two $\lambda/4$ CPW resonators with wide strip width or low impedance.

Figure 5.19: Layout of 2-stage J-coupling CPW filter

In the two sides, the 50Ω CPW feed lines are constructed as well. The higher-impedance transformer with the strip width of 8mil has the impedance of about 110Ω . The resonator is formed with the strip width of 90mil that corresponds to the impedance of about 38Ω . The uniform $\lambda/4$ CPW resonators are designed with the length of 220mil in relation to the central operating frequency of 5.0GHz. Due to the very short length of the strip conductor in the lumped inductive element inductance, the relevant inductive coupling degree can not be high enough to match the strong coupling degree of the transformer based

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structures if the good passband performance is requested. To circumvent this problem, the two narrow slots are etched out at the two sides of the central strip conductor such that the concerned inductive coupling can be largely increased as the equivalent narrow strip length is stretched. Figure 5.20 shows the modified 2-stage filter layout.

Figure 5.20: Layout of 2-stage J-coupling filter with modified resonators

The strip and slot widths in the central inductive element are set as 5mil. The slot length (t) is gradually lengthened from 0 to 20mil and then 40mil. Figure 5.21 shows the simulated frequency responses of the reflection constants in these three cases. At $t = 0$, the filter with the layout shown in Figure 5.19 suffers from very low return loss with one minimum peak. As explained above, it is basically caused by the too weak coupling degree at the center. As t increases, the coupling degree seems to become tightened. As such, the minimum peak at $t = 20$ mil is significantly reduced below -40dB. At $t = 40$ mil, the two transmission poles or two minimum peaks appear in the passband of this filter and the return loss within the passband is higher than 15dB. As such, the better passband performances are achieved with the use of the improved inductive coupling structure at center. In addition, the central operating frequency finds to be slightly shifted to the low frequency due to the physical reason discussed in Section 5.2 in this chapter.

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Figure 5.21: Frequency response of modified K-coupling element

After careful adjustment of this slot length (t), the optimized performance in the dominant passband is achieved in theory and the relevant circuit sample is then fabricated for measurement. The filter is designed with the Momentum simulator on the substrate of RT/Duroid 6010 with $\epsilon_r = 10.8$ and $h = 25$ mil.

Figure 5.22 shows the simulated and measured results in the narrow and wide frequency ranges. Both them agree well with each other over the frequency range, achieving the dominant passband with the bandwidth of about 28%, the two transmission poles and the return loss higher than 15dB. The lowest harmonic passband appears around 14.0GHz that is much higher than 5.0GHz of its dominant passband, thus exhibiting us that the upper stopband is also largely widened as the other CPW filters discussed above.

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(a) Narrow frequency range

(b) Wide frequency range

Figure 5.22: Frequency response of 2-stage J-coupling CPW filter

CHAPTER 6

CONCLUSION AND FUTURE WORK

6.1 Conclusion

Coplanar waveguide (CPW) has gained a wide application in the exploitation of uniplanar microwave integrated circuits because of its most attractive feature over the microstrip line counterpart, i.e., easy formulation of shunt and series circuit elements and easy mounting of external circuit blocks in shunt. During the period of this M. Eng. research work, extensive work has been carried out to design of various CPW bandpass filters with miniaturized size and harmonic suppression in addition to characterization of improved CPW periodic structures with inductive loading.

CPW resonator not only occupies an extremely large portion of the overall dimension in the CPW filter design and also dominate the eventual stopband performance in terms of its existed multiple resonance. The traditional uniform $\lambda/4$ CPW resonator with open- or short-ends at two terminals causes the first harmonic resonance to be happened at the frequency (f_1) twice its corresponding fundamental resonant frequency (f_0). To address this issue, the CPW stepped-impedance resonator (SIR) is initially presented to raise the lowest harmonic to the three times of its fundamental counterpart, i.e., $f_1=3f_0$. Furthermore, a number of modified $\lambda/2$ CPW resonators are proposed and characterized

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to significantly increase the lowest harmonic beyond $3f_0$. As such, a very wide upper stopband is realized as exhibited in simulation and experiment.

Next, in order to effectively tighten the coupling degree for filters with relatively wide passband and low insertion loss, several distributed coupling structures are proposed. Their coupling behaviors are characterized in terms of the relevant equivalent circuits and/or simulated results. In addition to the inductive overlapped coupling element, the $\lambda/4$ CPW impedance transformer with low or high characteristic impedances is modeled and used as the simple inductive or capacitive coupling elements with enhanced coupling degree.

With the use of those resonators and coupling structures, a number of novel CPW bandpass filters are designed, implemented and measured to demonstrate their attractive features such as low insertion loss, relatively wide passband, extremely wide stopband, miniaturized size and so on. The optimization design of these filters are carried out by properly adjusting several dimensions such as resonator length, coupled-end width or spacing, overlapped-coupling length and strip/slot width of impedance transformer via numerical simulation on the fullwave Momentum simulator. The predicted results of those filters with two and four resonators are confirmed by measurement of fabricated circuit samples over a wide frequency range.

6.2 Recommendations for future work

In this M. Eng. research project, all the design procedure is eventually carried out relying on the fullwave simulation. It usually takes much long CPU time and requires large CPU

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memory, as I experienced here, especially optimization of multiple-stage bandpass filters. It is preferred for one to extract the circuit models of each discontinuity elements such as step discontinuity in the SIR resonator from the fullwave simulation [16] and further develop the simple unified synthesis approach [28] based on the J- or K-inverter parameters and the susceptance or reactance slope of the $\lambda/4$ and $\lambda/2$ CPW resonators.

In order to further improve the passband performance with sharp rejection and wide/deep upper stopband, the multiple-stage CPW filters should be constituted and implemented. More interestingly, the cross-coupling technique was for the first time systematically reported in [29] and is very recently employed to explore a novel class of CPW filters with the two transmission zeros below and above the dominant passband [30]. This technique can also be used in the future to make up the CPW filters using the proposed CPW coupling structures and improved SIR resonators.

In addition, the periodic CPW structures with slow-wave and bandstop properties, as discussed in Chapter 2, may be used to construct the improved CPW resonators with miniaturized size and wide stopband. Meanwhile, this periodic CPW with increased impedance as derived in Chapter 2 may be alternatively employed to build up the miniaturized $\lambda/4$ transformer based capacitive coupling structures with preferably reduced coupling degree for design of filters discussed in Section 5.7.

PUBLICATIONS

PUBLICATIONS

Journal Papers

[1] **H. Shi** and L. Zhu, "High-impedance transformer for wideband $\lambda/4$ CPW bandpass filters", *IEICE Electronic Express*, vol.1, no.7, pp.198-203, July 2004.

[2] L. Zhu, **H. Shi** and W. Menzel, "Coupling behaviours of quarter-wavelength impedance transformers for wideband CPW bandpass filters," *IEEE Microwave Wireless Components Lett.*, vol.15, no.1, pp.13-15, Jan. 2005.

[3] L. Zhu and **H. Shi**, "Frequency-dependent guided-wave characteristics of periodically series-capacitive loaded coplanar waveguides", to appear in *Microwave Optical Technology Lett.*, 2005.

Conference Papers

[4] L. Zhu, **H. Shi** and T. Yakabe, "Characterization of periodic CPW structures with inductive loading using cascaded transmission line network: guided-wave propagation", *2003 Asia-Pacific Microwave Conf. Proc.*, pp. 1458-1461, Seoul, Korea, Nov. 2003.

REFERENCES

REFERENCES

- [1] D. F. Williams and S. E. Schwarz, "Design and performance of coplanar waveguide bandpass filters" *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, no.7, pp. 558-566, July 1983.
- [2] D. M. Pozar, "Microwave Engineering", John Wiley & Sons, 1998, 2nd Ed.
- [3] K. C. Gupta, R. Garg and I. J. Bahl, "Microstrip Lines and Slotlines", Artech London, 2nd Edition, 1996.
- [4] C. P. Wen, "Coplanar waveguide: a surface strip transmission Line suitable for non-reciprocal gyromagnetic devices applications," *IEEE Trans. Microwave Theory Tech.*, Vol. MTT-17, no.12, pp. 1087-1090, Dec. 1969.
- [5] R. N. Simons, "Coplanar Waveguide Circuits, Components, and Systems", John Wiley & Sons, 2001.
- [6] K. A. Everard and K. M. Cheng, "High performance direct coupled bandpass filters on coplanar waveguide", *IEEE Trans, Microwave Theory Tech.* vol. MTT-41, no. 9, pp. 1568-1573, Sept. 1993.
- [7] R. E. Collin, "Foundations for Microwave Engineering", McGraw-Hill, Inc., 1992. 2nd Ed.

REFERENCES

- [8] J. P. Linner and E. L. Kollberg, "CAD models for shielded multilayered CPW", *IEEE Trans, Microwave Theory Tech.*, vol. MTT-43, no. 4, pp. 772-779, Apr. 1995.
- [9] T. Tsujichi, H. Matsumoto and T. Nishikawa, "A miniaturized end-coupled bandpass filters", *IEEE Trans. Microwave Theory Tech.*, vol. MTT-41, no. 9, pp. 1568-1573, Sept. 1993.
- [10] T. Paillot, P. Blondy, D. Cros and P. Y. Guillon, "A novel compact filter", 2002 *IEEE MTT-S Int. Microwave Symp. Dig.*, pp. 1793-1796, June 2002.
- [11] L. Zhu, "Slow-wave and bandgap transmission characteristic of finite periodic coplanar waveguide", 2002 *Asia-Pacific Microwave Conf. Proc.*, pp.1153-1156, Kyoto, Japan, Nov. 2002.
- [12] A. Gorur, G. Karpus and M. Alkan, "Characteristics of periodically loaded CPW structure", *IEEE Microwave Guided Wave Lett.*, vol. 8, no. 8, pp. 278-280, Aug. 1998.
- [13] J. Sor, Y. Qian and T. Itoh, "Miniature low-loss CPW periodic structure for filter applications", *IEEE Trans, Microwave Theory Tech.*, vol. MTT-49, no.12, pp. 2336-2341, Dec. 2001.
- [14] L. Zhu, H. Shi and T. Yakabe, "Characterization of periodic CPW structures with inductive loading using cascaded transmission line network: guided-wave propagation", 2003 *Asia-Pacific Microwave Conf. Proc.*, pp. 1458-1461, Seoul, Korea, Nov. 2003.

REFERENCES

- [15] L. Zhu, "Guided-Wave characteristics of period coplanar waveguides with inductive loading – unit-length transmission parameters", *IEEE. Trans, Microwave Theory Tech*, vol. MTT-51, no. 10, pp. 2133-2138, Oct. 2003.
- [16] L. Zhu, "Realistic equivalent circuit model of coplanar waveguide open circuit: lossy shunt resonator network," *IEEE Microwave wireless components Lett.*, vol. 12, no. 5, pp. 175-177, May 2002.
- [17] R. Fu, "Novel design of planar printed antennas with enhanced performance", Thesis of Master Engineering, Nanyang Technological University, 2004.
- [18] T. Tsujichi, H. Matsumoto and T. Nishikawa, "A miniaturized end-coupled bandpass filter using $\lambda/4$ hair-pin coplanar resonators," *1998 IEEE MTT-S Int. Microwave Symp. Dig.*, pp.829-302, June 1998.
- [19] M. Makimoto, and S. Yamashita, "Bandpass filters using parallel coupled stripline stepped impedance resonators," *IEEE Trans. Microwave Theory Tech.*, vol. 28, no. 12, pp. 1413-1417, Dec. 1980.
- [20] A. Sanada, H. Takehara, T. Yamamoto and I. Awai, " $\lambda/4$ step impedance resonator bandpass filters fabricated on coplanar waveguide", *2002 IEEE MTT-S Int. Microwave Symposium Dig.*, pp.385-388, June 2002.

REFERENCES

- [21] J. Gao and L. Zhu, "Investigation on asymmetric parallel-coupled CPW for $\lambda/4$ bandpass filters with broad rejection band," *IEICE Electronic Express*, vol.1, no.1, pp.1-6, Apr. 2004.
- [22] H. Shi and L. Zhu, "High-impedance transformer for wideband $\lambda/4$ CPW bandpass filters", *IEICE Electronic Express*, vol.1, no.7, pp.198-203, July 2004.
- [23] L. Zhu, H. Shi and W. Menzel, "Coupling behaviors of quarter-wavelength impedance transformers for wideband CPW bandpass filters," *IEEE Microwave Wireless Components Lett.*, vol.15, no.1, pp.13-15, Jan. 2005.
- [24] X. Wu and I. Awai, "Numerical and experimental investigation on quality factors of coplanar waveguide resonators," *IEICE Trans on Electronics*, vol.E83-C, no.12, pp. 1923-1929, Dec. 2000.
- [25] T. M. Weller "edge-coupled coplanar waveguide bandpass filter design", *IEEE Trans. Microwave Theory Tech.*, vol. MTT-48, no. 12, pp. 2453-2458, Dec. 2000.
- [26] W. Menzel, W. Schwab and G. Strause, "Investigation of Coupling Structures for Coplanar Bandpass Filters", *2002 IEEE MTT-S Int. Microwave Symposium Dig.*, pp. 1407-1410, June 1995.
- [27] L. Zhu and T. Yakabe, "Fullwave MoM-SOC technique for extraction of equivalent circuit models coplanar waveguide discontinuities: CPW- and CSL-mode cases," *IEICE Trans. Electron.*, vol.E86-C, no.11, pp.2292-2299, Nov. 2003.

REFERENCES

- [28] G. Matthaei, L. Young and E. M. T. Jones, "Microwave Filters, Impedance-Matching Networks, and Coupling Structures," *Artech House*, Norwood, 1980.
- [29] J.-S. Hong and M. J. Lancaster, "Couplings of microstrip square open-loop resonators for cross-coupled planar microwave filters," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-44, no. 12, pp. 2099-2109, Dec. 1996.
- [30] J. Zhou, M. J. Lancaster and F. Huang, "Coplanar quarter-wavelength quasi-elliptic filters without board-wire bridges," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-52, no. 4, pp. 1150-1156, Apr. 2004.

APPENDIX A

Appendix A:

Conformal mapping Techniques

As described in chapter 2, we can divide the whole CPW section into two homogeneous partial regions: the substrate region ($\epsilon_o \epsilon_r$) and the absence of the substrate region (ϵ_o).

As summarized in [5], the total capacitance C_{cpw} of the CPW transmission line is equal to the sum of the partial capacitances in the two partial regions: C_l in the substrate layer and C_{air} in the absence of the substrate layer. Thus, we can derive $C_{cpw} = C_l + C_{air}$.

These two partial capacitances, C_l and C_{air} , can be expressed as below:

$$C_l = 2\epsilon_o (\epsilon_r - 1) \frac{K(k_1)}{K(k_1')} \quad (A-1)$$

$$C_{air} = 4\epsilon_o \frac{K(k_0)}{K(k_0')} \quad (A-2)$$

where the modulus of the complete elliptic integrals function $K(x)$

$$k_1 = \frac{\sinh(\pi a / 2h)}{\sinh(\pi b / 2h)} \text{ and } k_1' = \sqrt{1 - k_1^2}$$

$$k_0 = S / (2W + S) = a / b$$

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Thus the total capacitance is given by:

$$C_{CPW} = 2\varepsilon_0(\varepsilon_r - 1) \frac{K(k_1)}{K(k_1')} + 4\varepsilon_0 \frac{K(k_o)}{K(k_o')} \quad (A-3)$$

The complete elliptic integrals function K is defined as below [5, 8]:

Assume $m = k^2$ and $m_1 = 1 - k^2$

$$\frac{K(k)}{K(k')} = -\pi \left\{ \ln\left[\frac{m}{16}\right] + 8\left(\frac{m}{16}\right)^2 + 84\left(\frac{m}{16}\right)^3 + 992\left(\frac{m}{16}\right)^4 \right\}^{-1} \text{ for } 0 < k \leq 1/\sqrt{2} \quad (A-4)$$

This can be further approximately expressed as $-\pi \left\{ \ln\left(2 * \frac{1 + \sqrt{k'}}{1 - \sqrt{k'}}\right) \right\}^{-1}$

$$\frac{K(k)}{K(k')} = -\frac{1}{\pi} \left\{ \ln\left[\frac{m_1}{16}\right] + 8\left(\frac{m_1}{16}\right)^2 + 84\left(\frac{m_1}{16}\right)^3 + 992\left(\frac{m_1}{16}\right)^4 + \dots \right\}^{-1} = \frac{1}{\pi} \left\{ \ln\left(2 * \frac{1 + \sqrt{k}}{1 - \sqrt{k}}\right) \right\} \text{ for } 1 > k \geq 1/\sqrt{2} \quad (A-5)$$

Thus, the effective dielectric constant and characteristic impedance of CPW are derived

$$\varepsilon_{eff} = \frac{C_{CPW}}{C_{air}} = 1 + \frac{\varepsilon_r - 1}{2} * \frac{K(k_1)}{K(k_1')} \frac{K(k_o')}{K(k_o)} \quad (A-6)$$

$$Z_o = \frac{1}{cC_{air}\sqrt{\varepsilon_{eff}}} = \frac{30\pi}{\sqrt{\varepsilon_{eff}}} \frac{K(k_o')}{K(k_o)} \quad (A-7)$$