Fullwave modeling of coplanar-waveguide discontinuities for applications in filter design and metamaterial characterization

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**Fullwave Modeling of Coplanar-waveguide
Discontinuities for Applications in Filter Design
and Metamaterial Characterization**

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School of Electrical & Electronic Engineering

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

2006

Statement of Originality

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

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Date

.....

GAO JING

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Summary

Coplanar waveguide (CPW) has been finding wide applications in microwave integrated circuit (MIC) and monolithic microwave integrated circuit (MMIC) due to several advantages it offers over microstrip lines. Typical discontinuities characteristics are an integral part of practical circuit design. Therefore, a good understanding of their circuit behavior is essential for design success. In the past few decades, great effort has been made to accurate characterization of a variety of CPW circuits and discontinuities by developing different advanced fullwave approaches, of which the method of moments (MoM) has been considered as the most efficient and powerful algorithm. In this way, several commercial MoM simulators have been explored in the recent years for direct simulation and optimization of a large number of CPW circuit blocks with complicated configurations. However, such a direct approach requires the intensive CPU time and huge memory thus preventing one from physical insight into the operation principle of the whole CPW circuit block. Therefore, it is preferred to formulate the equivalent circuit models for many CPW elements such that the optimization can be executed via resultant network topology.

The main goal of this thesis is to investigate various CPW discontinuities for CPW-based filter design and metamaterial characterization. To accomplish this task, the hybrid fullwave approach based on the method of moments (MoM) and short-open calibration (SOC) procedure, namely, MoM-SOC technique is firstly employed via two-port CPW discontinuity to extract equivalent circuit network parameters, i.e., Y- or

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Z- matrix. Accurate equivalent circuit models of several types of coplanar waveguide (CPW) coupling elements are then presented. Afterwards, J- or K-inverter parameters versus frequency are derived to be utilized together for compact quarter-wavelength CPW bandpass filter design via simple cascaded network topology. Our optimized results are reasonably validated by ADS-simulation and experiment, both them exhibit good passband behavior with the broadened dominant passband, ultra-broad stopband and harmonic suppression.

Further, two novel ultra-wideband (UWB: 3.1~10.6GHz) bandpass filters on coplanar waveguide (CPW) are presented, designed and implemented. They are formed by open-ended/short-ended non-uniform or multiple-mode resonators and CPW parallel-coupled capacitor and inductor with tight coupling strength. As a result, two types of five-pole UWB bandpass filters with only one full-wavelength are constituted. The optimized results are confirmed by a full-wave simulation over the whole filter layout, two UWB filter samples are fabricated to demonstrate the actual UWB passband behavior with the bandwidth (110%) as well as insertion loss (<1dB) and return loss (>10dB).

Another type of work is characterization of 1-D metamaterial based on MoM-SOC method. Actually, such newly materials can be constructed by transmission lines with periodically loaded series capacitor and shunt inductor. Two effective per-unit-length transmission parameters, i.e., complex propagation constant and characteristic impedance, of this periodic structure, are demonstrated aiming at the forward- and backward-wave propagation properties in several distinctive left-handed (LH) and right-handed (RH) stopbands and passbands. Moreover, coplanar waveguide (CPW)

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metamaterials with varied LH and RH passbands are studied in terms of per-unit-length transmission parameters and magnitude/phase of transmission coefficient, respectively. The two CPW metamaterials are formed by simultaneously loading the lumped or distributed CPW series-capacitive and shunt-inductive elements in periodic intervals. It is further exhibited that the stopband between the LH and RH passbands can be wholly eliminated in the distributed CPW case. Finite-extent CPW metamaterials with 50Ω feed lines are also modeled in term of S parameters via simple cascaded transmission line theorem. Predicted magnitude and phase of transmission coefficient are confirmed by experiment, showing the distinctive backward- and forward-wave performances in the LH and RH passbands.

The software using in this thesis is developed in house based on the Matlab and Fortran77 platform, the commercial software for the verification of predicted results is Advance Design System (ADS 2003) in Agilent.

Abbreviations

Abbreviations

CPW	coplanar waveguide
MICs	microwave integrated circuits
MMICs	monolithic microwave integrated circuits
TEM	transverse electro magnetic
EBG	electromagnetic bandgap
MMR	multiple mode resonator
UWB	ultra wideband
LH	left-handed
RH	right-handed
MoM	method of moments
SOC	short open calibration
PEC	perfect electric conductor
MFIE	magnetic field integral equation
PWS	piecewise sinusoidal
1-D	one dimensional
2-D	two dimensional
3-D	three dimensional
CAD	computer-aided design
CPU	central processing unit
DUT	device under test
TRL	thru-reflection-line

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Chapter 1

Introduction

1.1 Motivation

A coplanar waveguide (CPW) fabricated on a dielectric substrate was first demonstrated by C. P. Wen [1] in 1969. Since then, tremendous progress has been made to development of various CPW based microwave integrated circuits (MICs) as well as monolithic microwave integrated circuits (MMICs) [2-5] .

A conventional CPW on a finite length substrate consists of a center strip conductor with width w and semi-infinite ground planes on either side as shown in Figure 1.1. Such kind of structure can support quasi-TEM mode of propagation. It can offer many advantages over conventional microstrip line. These include the ease in connecting shunt lumped elements without using via holes, and the low dispersion of both the propagation constant and the characteristic impedance for the even CPW mode [6]. Also, the propagation constant and the characteristic impedance of CPW can be adjusted by changing the slot width to strip width ratio [7]. Thus its propagation characteristic is not so sensitive to the substrate thickness. Also there are several

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advantages in manufacturing. First, it makes use of automatic pick-and-place and bond assembly equipment for surface-mount component placement and interconnection of components [7]. Second, it allows computer controlled on-wafer measurement techniques for device and circuit characterization up to several ten GHz [8].

Generally speaking, three types of CPW transmission line can be classified as conventional CPW, conductor backed CPW, micromachined CPW. In a conventional CPW, the ground planes are semi-infinite extent on either side. However, in a practical circuit the ground planes are made of finite extent. Only this conventional CPW with finite substrate will be considered in this thesis. The conductor-backed CPW has an additional ground plane at the bottom surface of the substrate. This lower ground plane not only provides mechanical support to the substrate but also acts as a heat sink for circuits with active devices. But, it always suffers from effects of higher-order parallel-plate-waveguide mode. The micromachined CPWs have two types, namely, the microshield line [9] and the CPW suspended by a silicon dioxide membrane above a micromachined groove. The advantages of the microshield line are its extremely wide bandwidth, minimal dispersion and zero dielectric loss.

Typical applications of CPW include millimeter-wave amplifiers [10-12], active combiners [13], frequency doublers [14], mixers [15], filters [16-20], micro-electromechanical systems switches [21], thin film high-temperature superconducting tunable circuits and components [22], photonic bandgap structures or electromagnetic-bandgap (EBG) structure [23, 24], printed antennas [25] and so on.

Figure 1.1 Schematic of a coplanar waveguide on a dielectric substrate of finite thickness

As one of the family of transmission line structure, the analysis methods for a CPW to determine the characteristic impedance and dielectric constant can be divided into two main groups: quasi-static approach and fullwave analysis. In the first group, which consists of quasi-static method[1], modified conformal transformation method, finite difference method, integral equation method, variational method in frequency to time domain, boundary element method, the nature of the mode of propagation is considered as pure transverse electro magnetic (TEM) and the characteristic impedance is calculated from electrostatic capacitance. This is only suitable for lower frequency. The full-wave analysis is considered as the most accurate method in taking into account the frequency dispersion of characteristic impedances and dielectric constant of different structures. The methods in the second group include Galerkin's method in spectral domain [26-32], integral equation method [33], and finite difference time-domain method [34, 35].

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With great development of MMIC, CPW discontinuities play more and more important role in planar integrated circuit design [36]. The most commonly used CPW discontinuities include open-end [37], short end, series gap [1], step-impedance [38], spiral inductors [39]/capacitors and interdigital capacitors [40]. Also coplanar waveguide series as well as shunt stubs are essential for constructing filter circuits using capacitive and inductive elements [41]; the values of these capacitances and inductances are of a function of stub length, geometry as well as frequency [42, 43]. However, one of the main obstacles for CPW is the unwanted conversion from the dominant CPW (even) mode to the coupled slotline (odd) mode in asymmetric discontinuities, like bends and T-junctions [44], which are unavoidable in modern circuits where packing density increases with the complexity of the design. So all the modeling discontinuities and circuits considered in this thesis are horizontally symmetric with central strip.

Over the past few years, many numerical methods have been developed to analyze different planar structures, of them, the integral equation based method of moments is considered as very accurate and efficient tool for modeling of electromagnetic propagation, coupling and discontinuities. But this method exhibits a distinct difficulty in the implementation of adequate excitation mechanisms. Three commonly used approaches include traveling-wave excitation mechanism [45], impressed-field excitation with the concept of δ -gap generator [46], matched load simulation [47] developed for the de-embedding of port discontinuities in open printed circuits. In this

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way, a number of commercial MoM simulators [48-50] have been explored in recent years for direct simulation and optimization of a large number of CPW circuit blocks with complicated configurations. However, regardless of any of the excited-field methods, the parasitic effects brought by the introduced source port should be removed when considering electrically small discontinuities. This problem has been successfully solved by a newly proposed method of moments and short-open calibration (MoM-SOC) technique [51, 52]. It also has been proved as the most accurate and efficient way to de-embed the two port planar circuit elements [53]. However, there still exists a big gap between numerical electromagnetic analysis and accurate circuit model for microwave integrated circuit design. In this aspect, our developed MoM-SOC technique is extended in this work to model several types of CPW discontinuities and extract their equivalent circuit parameters such that the optimization and design criteria can be executed via resultant network topology for the filter design in an efficient way.

Planar transmission lines with periodically loaded obstacles have been widely studied in the past several decades to construct the size-compact and frequency selective devices due to their slow-wave behaviors and guided-wave passband and stopband characteristics [54]. Besides various planar electromagnetic bandgap (EBG) structures with bandstop behavior [55-59], more and more research interests are concentrated on a new class of planar transmission line structures that can be operated as 1-D [60] or 2-D [61] artificial medium or metamaterial with simultaneously negative permittivity (ϵ) and permeability (μ). The concept of this material were originally theoretically proposed by Veselago in 1968 [62]. In 2001, these metamaterials based on split ring

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resonators and continuous wires were successfully implemented to experimentally exhibit the existence of negative effective refractive index in a certain frequency passband [63, 64]. In addition to these bulky 3-D structures, 1-D and 2-D transmission line based metamaterials were constituted by periodically interrupting the uniform microstrip line with the capacitor and inductor chip in series and shunt formats, thus exhibiting some basic guided-wave propagation phenomenon in the passbands such as negative phase constant or anti-paralleled phase and group velocities [65-67] using traditional transmission line theorem [54]. However, to our knowledge, little effort has been reported today to characterize the fully integrated and inhomogeneous transmission line metamaterials in terms of effective per-unit-length series capacitance and shunt inductance of the 1-D composite metamaterial with infinite extension. Thus, in this thesis, two effective per-unit-length transmission line parameters, i.e., characteristic impedance and propagation constant are derived through CPW periodic structure and transmission line theorem to demonstrate fundamental forward and backward wave propagation properties.

1.2 Objectives

According to the problems mentioned above, the goal of this thesis is to investigate the circuit parameters of different CPW discontinuities using MoM-SOC technique. These circuit parameters are then utilized for efficient and accurate design of CPW filters and metamaterials analysis. To fulfill these goals, the following tasks should be envisaged.

1. Extensive literature reviews are made about different numerical modeling

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techniques for various CPW discontinuities and circuits.

2. Fullwave MoM-SOC technique established for parametric extraction of various CPW discontinuities and circuit elements.
3. CPW capacitive and inductive coupling elements are characterized in terms of J- and K- inverters for optimization design of size-miniaturized and harmonic-suppressed CPW bandpass filters.
4. Open- and short-ended CPW multiple-mode resonators are proposed and utilized for innovative design of a new class of UWB bandpass filters.
5. Various CPW metamaterials with periodical loading of series-capacitor and shunt-inductor are investigated in terms of per-unit-length transmission parameters.
6. Guided-wave characteristics of 1-D CPW metamaterials with varied or tunable left-handed and right-handed passbands are further studied.
7. Experimental verification is evidently made for the finite-extent CPW metamaterial blocks and CPW bandpass filters.

1.3 Major Contribution of This Thesis

In this thesis, several original contributions have been achieved and the relevant outcomes have led to publication and submission of several publications. The major contributions are listed below:

A fullwave and hybrid approach based on method of moments (MoM) and short-open

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calibration (SOC), namely, (MoM-SOC) technique, is developed and used to numerically de-embed the equivalent circuit models of various CPW discontinuities. They include the lumped/distributed capacitive/inductive coupling elements and asymmetric paralleled-coupled or overlapped inductive coupling elements and enhanced open stub, in addition to the frequency dispersive characteristic impedance and effective dielectric constant of uniform CPW transmission line.

Compact quarter-wave resonator CPW bandpass filters are designed via simple cascaded network topology. For this purpose, the impedance and admittance inverters of CPW elements are obtained from their corresponding equivalent T- or π - type networks. To generate a transmission zero, an overlapped coupling structure is constructed by differentiating the coupled spaces and overlapped lengths of two coupling element. As a result, a class of improved two-stage and four-stage quarter-wavelength bandpass filters with 2nd harmonic suppression are realized. Both predicted results are validated by the fullwave simulation on the Agilent ADS software and experiment of the fabricated circuit samples.

A new class of ultra wide band (UWB) bandpass filters (3.1GHz-10.6GHz) on CPW is presented by using multiple mode resonators (MMR) and distributed CPW element with tight coupling. The proposed MMR is utilized to generate three resonant frequencies at lower-end, central and upper end of UWB band while the tight coupling element with two impedance steps generates two additional poles located at two sides of the central frequency of UWB passband, i.e., 6.85GHz, to improve the insertion loss performances. Two UWB filter samples are designed and fabricated to demonstrate the

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actual UWB behaviors.

Transmission line metamaterials on CPW with series-capacitive and shunt-inductive interruption in periodical intervals are thoroughly characterized using a single cascaded transmission line network with de-embedded circuit parameters. The complex characteristic impedance and propagation constant of infinite-extent metamaterial are at first derived to demonstrate the forward and backward guided-wave propagation characteristics. Then, the transmission performances of the finite-cell metamaterial blocks are initially characterized under the assumption of perfect impedance matching cases. Two 9-unit-cell metamaterial circuits fed by two 50Ω feed lines are implemented for experimental verification on the derived per-unit length parameters.

Extensive investigation and discussion on CPW 1-D metamaterials with varied left-hand (LH) and right-hand (RH) passbands are further made in terms of per-unit-length transmission parameters and magnitude/phase of transmission coefficient, respectively. Various CPW metamaterials are constructed by simultaneously loading the lumped or distributed CPW series-capacitive and shunt-inductive elements in periodic intervals. Effective characteristic impedance and propagation constant of these infinite-extent metamaterials are extracted to give a physical insight into the guided-wave propagation characteristics in the LH/RH passbands. It is exhibited that the stopband between the LH and RH passbands can be wholly eliminated in the distributed CPW case. The two finite-extent CPW metamaterials with 50Ω feed lines are then modeled via simple cascaded transmission line theorem. Predicted magnitude and phase of transmission

coefficient are confirmed by experiment, showing the distinctive backward- and forward-wave performances in the LH and RH passbands.

1.4 Organization of This Thesis

The reminder of this thesis is organized as follows. Chapter 2 describes the MoM-SOC de-embedding technique to calibrate out the port discontinuity effects introduced by impressed field. Several two-port CPW discontinuities are modeled in terms of a two-port network, i.e., Z- or Y- matrix.

In chapter 3, the modeling method based on dyadic green's function and magnetic integral equations are described to formulate a fullwave method of moments (MoM). Combined with the SOC technique, this MoM technique is utilized to characterize several CPW discontinuities such as open-end inductive/capacitive element, distributed enhanced capacitive/inductive coupling elements and asymmetrical overlapped inductive coupling elements.

In chapter 4, accurate circuit models of CPW coupling elements are presented to design quarter-wavelength filters in terms of impedance or admittance inverters (J or K inverter) parameters. An asymmetric overlapped CPW coupling structure is originally characterized as an equivalent K-inverter network. By adjusting the coupling length and coupled spaces, transmission zeros can be allocated at two unequal locations around 2nd spurious harmonic. Finally, two and four stage filters with good harmonic

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suppression are designed and fabricated to put this idea into reality.

In chapter 5, two novel UWB bandpass filters on coplanar waveguide (CPW) are proposed, designed and implemented. Open ended or short ended CPW multiple-mode resonator (MMR) are characterized to allocate its first three resonant modes occurring near the lower-end, center and higher-end of the targeted UWB passband. Then, a CPW interdigital capacitor or paralleled-coupled inductor is investigated to excite two poles below and above the UWB's center or 6.85GHz. Thus, five-pole CPW wideband filters with the UWB passband are designed on a basis of simple equivalent transmission line network. Our network-predicted performances are confirmed by fullwave simulation of an overall filter layout, two CPW filters are fabricated to show the actual UWB passband behaviours with insertion loss $< 0.5\text{dB}$, return loss $> 10\text{dB}$ and group delay variation $< 0.23\text{ns}$ over the entire UWB passband.

In Chapter 6, we demonstrate 1-D transmission line metamaterials with series-capacitive and shunt-inductive interrupting in periodical intervals. Two effective per-unit-length transmission line parameters, i.e., complex propagation constant and characteristic impedance, are analyzed to investigate the forward- and backward-wave propagation properties in several distinctive left- and right- handed stopbands and passbands. These parameters can be further converted to effective material physics characteristic, i.e., effective permeability and permittivity of concern. The predicted results and extracted parameters are validated by experiment of implementation of 9-unit cell CPW circuits.

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In chapter 7, CPW metamaterials with varied left-handed and right-handed passbands are studied via different periodic load of lumped and distributed series capacitor and shunt inductor. It is found that the LH and RH passbands can be linked together with seamless gap if those coupling elements are properly constructed. Finally, two 9-unit cell metamaterials with lumped and distributed elements are fabricated to verify the predicted transmission coefficients.

Finally, Chapter 8 is the conclusion of this thesis and the recommendations for future work.

Chapter 2

Fullwave MoM-SOC Technique

Abstract: In this chapter, the Method of Moments (MoM) algorithm and short-open calibration (SOC) technique are described through the full-wave modeling of the two-port CPW discontinuity.

2.1 Introduction

With development in monolithic and hybrid integrated circuit (MMIC) technology and the abundant availability of low-cost high-speed computers, considerable interests have been focused on directly developing efficient full-wave algorithms for modeling and design as well as optimization of high-frequency planar structures [46, 68, 69]. Over the last 20 years, a significant amount of work has been investigated toward field-theoretical analysis and modeling of a class of planar integrated circuits and discontinuities that are bounded or unbounded enclosed [70-73]. Within the framework of time-harmonic fields, both spatial [50, 70, 74] and spectral-domain [51, 71]

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techniques have been successfully implemented. Due to the surface discretization of the circuit instead of the volumetric discretization of the space, fullwave method of moments (MoM) is considered as the most efficient and accurate method compared to the differential-equation method [72, 75]. However, this method faces a distinct difficulty in the implementation of adequate excitation mechanisms [45, 46].

To my knowledge, this difficulty definitely occurs in the different excitation approaches. In [45], the traveling wave excitation was implemented, where the entire domain traveling-wave basis functions on the infinite transmission lines is introduced in the MoM algorithm. Although this technique preserves the correct physics of the circuit excitation, numerical difficulties associated with the evaluation of the semi-infinite reaction integrals for the transmission-line currents make it rather cumbersome for general-purpose multilayered multiport structure implementation. Another approach called impress-field excitation utilized the δ -gap generator and owes its name to the fact that the electric field is impressed at an infinitely small gap between the stub end of the transmission line and the perfect electric conductor (PEC) plane to provide the ground reference for the formed voltage generator [46, 69]. The third method called matched load approach, was developed for de-embedding of port discontinuities in open printed circuits [47]. Similarly to the standing wave excitation mechanism, this technique requires computation of the cross-sectional current distribution and the propagation constants of the quasi-TEM eigenmodes of the associated multiconductor transmission-line networks.

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Regardless of all these excitation mechanisms, the potential error terms or parasitic effect brought by the approximation of such MoM algorithms can not be negligible if an electrically small discontinuity such as gap coupling element is involved. Thanks to the recently proposed short-open calibrated method (SOC) in the δ – gap field-impression excitation in [52], a novel de-embedding technique has been developed as the most accurate and efficient way to model two-port or multiport [76] planar discontinuities without receiving the parasitic effects of the impressed source.

In this chapter, the short-open calibration (SOC) method is introduced based on the two-port CPW discontinuity. Magnetic field integral equation (MFIE) is first established to give matrix equation through basis and testing functions. Two calibration standards, called open-end and short-end are implemented to solve ABCD transfer matrix of the port or feed lines so as to finally extract the Y- or Z-matrix of the device under test (DUT).

2.2 Field-theoretical MoM

Among the field-theoretical-based algorithms developed to date, the method of moments (MoM) formulated in the spectral domain or space domain is considered as one of the most powerful tools for comprehensive treatment of various complex planar structures. A MoM-based technique is applied to the numerical handling of the multi-port circuit such that a systematic integral equation can be built up for the solution of

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the unknown current density.

Figure 2.1 The arrangement of basis functions and impressed current density

To begin with the following MoM analysis, we consider a two-port CPW discontinuity as shown in Figure 2.1. In order to model this two-port CPW discontinuity circuit, a magnetic-field integral equation (MFIE) governing the total magnetic current density ($\vec{M}$) can be established on the slotline surface that includes each uniform CPW feed line and the CPW discontinuity.

With the two distinct parts in mind, we can formulate the MEIE for a two-port circuit such that

$$\sum_{j=1}^2 [\vec{J}_j^{inc}(\vec{r}) + \hat{z} \times \iint_{line(j)} \overline{\overline{G}}_{HM}(\vec{r}, \vec{r}_s) \cdot (\vec{M}_j(\vec{r}_s)) d\vec{s}] + \hat{z} \times \iint_{circuit} \overline{\overline{G}}_{HM}(\vec{r}, \vec{r}_s) \cdot (\vec{M}(\vec{r}_s)) d\vec{s} = 0 \quad (2.1)$$

where $\vec{J}_j^{inc}(\vec{r})$ denotes the impressed electric current density along CPW feeding line and $\overline{\overline{G}}_{HM}(\vec{r}, \vec{r}_s)$ presents the dyadic Green's function and $\vec{M}(\vec{r}_s)$ is unknown magnetic current density.

2.2.1 Basis and Testing Functions

Under the framework of the MoM, the vector function of the magnetic current density over the slot surface is expanded into a set of subsectional basis functions as expressed in equation (2.2).

$$\begin{aligned}\vec{M}(\vec{r}_s) &= \sum_M I_m \vec{J}_m(x_s, y_s) \\ &= \sum_M (\hat{x} I_{xm} J_{xm} + \hat{y} I_{ym} J_{ym})\end{aligned}\quad (2.2)$$

where J_{xm} and J_{ym} are the basis functions regarding to x- and y- oriented current density while I_{xm} and I_{ym} are their corresponding unknown amplitude. These basis functions are composed of piecewise sinusoidal (PWS) functions in the longitudinal direction and pulse functions in the transversal direction as equations (2.3).

$$\begin{aligned}J_{ym}(x, y) &= \begin{cases} \frac{\sin k_s (\Delta y - |y - y_m|)}{\sin k_s \Delta y} & \left(\begin{array}{l} |y - y_m| \leq \Delta y \\ |x - x_m| \leq \Delta x/2 \end{array} \right) \\ 0 & \text{otherwise} \end{cases} \\ J_{xm}(x, y) &= \begin{cases} \frac{\sin k_s (\Delta x - |x - x_m|)}{\sin k_s \Delta x} & \left(\begin{array}{l} |x - x_m| \leq \Delta x \\ |y - y_m| \leq \Delta y/2 \end{array} \right) \\ 0 & \text{otherwise} \end{cases}\end{aligned}\quad (2.3)$$

The symbol k_s is an arbitrary constant which is selected as phase constant of dominant mode propagating along CPW feeding lines. The configuration of rectangular mesh of

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basis function is shown in Figure 2.1. Subsequently, a set of weighting functions are to be selected for solving the integral equation. Using the Galerkin's method, the testing functions are chosen the same as basis functions so as to transform the integral equation (2.1) to its related matrix equation.

$$\iint_{slot} \bar{J}_n^w(x, y) \cdot \left(\sum_{j=1}^2 [\bar{J}_j^{inc}(\vec{r})] \right) + \hat{z} \times \iint_{line(j)} \bar{G}_{HM}(\vec{r}, \vec{r}_s) \cdot (\bar{M}_j(\vec{r}_s)) ds + \hat{z} \times \iint_{circuit} \bar{G}_{HM}(\vec{r}, \vec{r}_s) \cdot (\bar{M}_j(\vec{r}_s)) ds = 0$$

where

$$\bar{J}_n^w(x, y) = \bar{J}_n(x, y) = \begin{cases} \hat{x} J_{xn}(x, y) \\ \hat{y} J_{yn}(x, y) \end{cases}$$

2.2.2 Matrix Equations

Fourier transform can lead to the formulation of spectral-domain integral equation as expressed in equation (2.4)

$$\begin{aligned} & \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \tilde{J}_n^{w*}(\alpha, \beta) \cdot \bar{G}_{HM}(\alpha, \beta) \cdot \left(\sum_{j=1}^2 \tilde{M}_j(\alpha, \beta) + \tilde{M}(\alpha, \beta) \right) d\alpha d\beta \\ & = - \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \tilde{J}_n^{w*}(\alpha, \beta) \sum_{j=1}^2 \tilde{J}_j^{inc}(\alpha, \beta) d\alpha d\beta \end{aligned} \quad (2.4)$$

Without loss of generality, the feed lines are assumed to be far away from each other so that the admittance matrix associated with each feed line can be separately formulated.

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In this way, the explicit network parameters of two-port CPW can be derived from the extraction of the matrix equation in (2.5)

$$\begin{bmatrix} Y_{11} & \cdots Y_{1m} \\ \vdots & \vdots \\ Y_{m1} & \cdots Y_{mm} \end{bmatrix} \begin{bmatrix} V_1 \\ \vdots \\ V_m \end{bmatrix} = \begin{bmatrix} I_1 \\ \vdots \\ I_m \end{bmatrix}$$

$$I_m = - \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \tilde{J}_m^w(\alpha, \beta) \cdot \tilde{J}^{inc}(\alpha, \beta) d\alpha d\beta$$

$$Y_{mn} = \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \tilde{J}_m^w(\alpha, \beta) \cdot \tilde{\tilde{G}}_{HM}(\alpha, \beta) \cdot \tilde{J}_n(\alpha, \beta) d\alpha d\beta \quad (2.5)$$

where

$$\tilde{J}_{xm}(\alpha, \beta) = \begin{cases} \frac{J_{xm}(\beta) \cdot e^{j\alpha x_m}}{\sin(k_s \Delta x) \cdot (\alpha^2 - k_s^2)} \cdot 2k_s [\cos(k_s \Delta x) - \cos(\alpha \Delta x)] \\ J_{xm}(\beta) = 2 \sin(\beta \Delta y / 2) / \beta \end{cases}$$

$$\tilde{J}_{ym}(\alpha, \beta) = \begin{cases} \frac{J_{ym}(\alpha) \cdot e^{j\beta y_m}}{\sin(k_s \Delta y) \cdot (\beta^2 - k_s^2)} \cdot 2k_s [\cos(k_s \Delta y) - \cos(\beta \Delta y)] \\ J_{ym}(\alpha) = 2 \sin(\alpha \Delta x / 2) / \alpha \end{cases} \quad (2.6)$$

By substituting (2.6) into (2.5), one can obtain

$$Y_{mn} = \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \left[\tilde{J}_{mx}^w(-\alpha, -\beta) \cdot \tilde{\tilde{G}}_{xx}(\alpha, \beta) \cdot \tilde{J}_{nx}(\alpha, \beta) + \tilde{J}_{mx}^w(-\alpha, -\beta) \cdot \tilde{\tilde{G}}_{xy}(\alpha, \beta) \cdot \tilde{J}_{ny}(\alpha, \beta) \right] d\alpha d\beta$$

$$+ \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} \left[\tilde{J}_{my}^w(-\alpha, -\beta) \cdot \tilde{\tilde{G}}_{yx}(\alpha, \beta) \cdot \tilde{J}_{nx}(\alpha, \beta) + \tilde{J}_{my}^w(-\alpha, -\beta) \cdot \tilde{\tilde{G}}_{yy}(\alpha, \beta) \cdot \tilde{J}_{ny}(\alpha, \beta) \right] d\alpha d\beta$$

2.3 SOC De-embedding Technique

SOC technique is directly accommodated in a full-wave method-of-moments (MoM) algorithm. The developed MoM algorithm is applied to the modeling of unbounded planar discontinuities that can be segmented into two distinct sections: static model of feed lines and dynamic model of circuit discontinuity. This method is used to remove unwanted parasitic brought by the approximation of the impressed source and also the problem of the resulting consistency between the two-dimensional and three-dimensional simulations.

2.3.1 Excitation Mechanisms

The MoM has been recognized as one of the most powerful candidates for accurate and efficient modeling of planar circuits. It has been well applied to solving bounded/unbounded problems in a comprehensive manner, namely, the consideration of realistic physical effects such as parasitic shielded resonance, space-wave radiation, and surface-wave leakage. Despite its success, this technique suffered from the difficulty of accurately terminating each feed line that is mathematically supposed to be in electrically infinite extent for the extraction of guided-wave characteristics in the modeling of opened/unbounded structures. In the middle of the 1980's, two workable excitation mechanisms, namely, traveling-wave excitation [45] and delta-gap voltage excitation [46] were initially proposed to remedy such a problem for modeling of a large variety of unbounded structures.

From the view of introducing impressed field, the MoM algorithms may be classified

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into two groups: indirect and direct algorithms. The indirect algorithm has been known as a very accurate tool in the characterization of electrical characteristics of a class of planar structures using simulated (standing-wave) field profiles along feed lines [45], [77], and [78]. However, due to the complex numerical de-embedding procedure in the extraction of the equivalent-circuit model, this technique is bound to be in the modeling of simplified structures such as one- and two-port discontinuities. The direct algorithm is derived from an effective hypothesis in that the impressed field is imposed in an approximate way at each feed line to formulate a deterministic algorithm. One of the most attractive advantages of this scheme is its ability to directly characterize a multiport discontinuity in terms of the calculated field/circuit parameters at each port reference plane without resort to additional calculations. Nevertheless, there is a common problem for these direct algorithms in that the impressed fields are theoretically impossible to excite ideal dominant modes.

The direct MoM algorithms have been successfully applied to the modeling of microstrip and CPW discontinuities having an electrically large volume such as a CPW resonator antenna and filter. Potential error terms (or parasitic effects) brought by the approximation of such algorithms are negligible and can hardly exceed a prescribed accuracy of numerical results. However, when an electrically small discontinuity is involved, such as a CPW coupling gap, the use of these algorithms may become problematic because these error terms could usually be in the same order as values of electrical parameters identified in the circuit model, e.g., the equivalent capacitance or inductance. This was documented in [79] and [80]. Strictly speaking, such a port

discontinuity should be modeled as a generalized circuit network involving all potential physical effects. On the other hand, there is a common practice in use of these algorithms that the dielectric substrate is usually considered to be electrically thin so that two-dimensional (2-D) precalculated characteristics of a quasi-TEM mode for each feed line are well valid. Obviously, the 2-D and three-dimensional (3-D) MoM calculations that are made through the use of two different and separate procedures may lead to some small difference that is, in general, dispersive visible unmatched difference between them. Even if it is very small and structure-dependent, this error cannot be automatically removed from the numerical model.

2.3.2 SOC Self-calibration Method

Analog to the well-established calibration procedures in the circuit experiments, a planar CPW discontinuity structure can be effectively divided into two separate parts: error boxes and a circuit required to be modeled, as shown in Figure2.2. The first part represents the potential error terms related to the port/transition discontinuities, i.e., error box $[X_1]$ and $[X_2]$, while the second part characterizes the realistic circuit properties. Using different types of standard calibration elements such as the thru-reflection-line (TRL) elements, these error boxes can be independently evaluated for the de-embedding of circuit box parameters. In the light of these considerations, an improved deterministic MoM algorithm using the impressed current excitation mechanism [77-79] has been successfully developed for field-theoretical characterization of CPW discontinuity circuits. In that work, specific interests were

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focused on the intrinsic physical differences between the feed lines and the circuit discontinuity, which form the two distinct sections of an entire unbounded structure. The modeling strategy is that the uniform feed lines are considered as static and excited separately by a set of local impressed current sources, while the circuit discontinuity is kept unbounded.

Figure 2.2 Equivalent-circuit representation of two-port CPW discontinuity calibration element for de-embedding technique

In this section, the short-open calibration (SOC) technique is described, which allows for a very accurate extraction of dynamic circuit parameters pertinent to an unbounded electrically small CPW discontinuity in an impressed current source-based MoM. Similar to the TRL calibration technique, such a SOC procedure is able to completely remove the potential errors or parasitic terms. These error terms are brought by the excitation model and inherent effects of each feed line out of the dynamic circuit model, which solely represents the region of the unbounded discontinuity. This SOC function is achieved effectively by use of the proposed two ideal line elements, namely, “open” and “short”. Dispersion and various parametric effects are considered. The physical

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models and circuit networks of the ideal “open” and “short” line elements realized by a pair of even and odd oriented electric-field excitations for the 3-D MoM algorithm are shown as Figure 2.3.

Now, the so-called “short” and “open” elements can actually be formulated in the MoM to accurately simulate the ideal open- and short-end according to the transmission-line theory. So it is realized by exciting a pair of even and odd impressed currents (J_{io} and J_{is} for electric current by equivalence). To handle the i^{th} port block, a uniform CPW transmission line is suitably selected with a length of twice the distance L_i , which is defined between the length of port and reference plane of the i^{th} feed line, as shown in Figure 2.3. Two impressed current sources are then imposed at the two sides of CPW feed line, which form the idealized short- and open-end, respectively.

Without loss of generality, a uniform line with a length of $2L_i$, whose two-port planes are subject to an excitation of two independent impressed electrical currents, can be dynamically formulated in the MoM, as described in equation (2.1).

To realize a “short” element in the proposed circuit model, a pair of impressed electric current having identical magnitude and orientation are used to excite at the two-port planes, thus,

$$\vec{J}_{i1} = \vec{J}_{i2} = -\hat{y}I_{is} \cdot \delta(s)$$

where s is the location of impressed current source.

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Figure 2.3 Physical models and circuit networks of ideal “open” and “short” line elements (a) Ideal “short” line element and its circuit model. (b) Ideal “open” line element and its circuit model.

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Similarly, a pair of impressed electric currents with identical magnitude and reverse orientation is used to excite the two-port planed in realizing an “open” element, thus,

$$\vec{J}_{i1} = -\vec{J}_{i2} = -\hat{y}I_{io} \cdot \delta(s)$$

In this open-end element, such a symmetrical plane represents a perfect magnetic wall in the proposed model, and its equivalent to an idealized open-end in the circuit network, as shown in Figure 2.3 (b). On the other hand, the electric voltage at the open-end can be expressed as a linear superposition of the magnitudes of the voltage computed by matrix equations in MoM (equation 2.5). The magnetic voltage distribution on symmetrical CPW discontinuity plane (T_i) is shown in Figure 2.4. The total magnitude of voltage is related to all the transverse meshes along the symmetrical plane as expressed in equation (2.7)

$$V'_{io} = \sum_M V_m \quad (2.7)$$

where V_m is the electric voltage along the symmetrical plane T_i .

Figure 2.4 Electric voltage distribution in CPW feed line

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Considering this 3-D precalculated ABCD matrix, the equivalent current and voltage I'_i and V'_i , defined at the two sides of the realistic circuit network characterizing the two-port discontinuity as shown in Figure 2.3, can be expressed in equation (2.8).

$$\begin{bmatrix} V' \\ I' \end{bmatrix} = [X_i] \cdot \begin{bmatrix} V \\ I \end{bmatrix} \quad (2.8)$$

ABCD matrix of X_i can be obtained by the calculated current and voltage in the “open” and “short” circuit elements, such that:

$$\begin{aligned} a_i V_{is} + b_i I_{is} &= 0 && \text{(short circuit)} \\ a_i V_{io} + b_i I_{io} &= V'_{io} && \text{(open circuit)} \\ c_i V_{io} + d_i I_{io} &= 0 && \text{(open circuit)} \\ a_i d_i - b_i c_i &= 1 && \text{(reciprocity)} \end{aligned} \quad (2.9)$$

To simplify the following description, all of the above equivalent voltages are, respectively, normalized by the impressed port currents I_{io} and I_{is} for the two kinds of excitation schemes:

$$\begin{aligned} \bar{V}_{io} &= V_{io} / I_{io} \\ \bar{V}_{is} &= V_{is} / I_{is} \\ \bar{V}'_{io} &= V'_{io} / I_{io} \end{aligned}$$

$$[x_i] = \begin{bmatrix} \bar{V}'_{io} / (\bar{V}_{io} - \bar{V}_{is}) & -\bar{V}'_{io} / (\bar{V}_{io} - \bar{V}_{is}) \\ -1/\bar{V}_{io} & \bar{V}_{io} / \bar{V}'_{io} \end{bmatrix} \quad (2.10)$$

In order to obtain the admittance matrix ($[Y]$) of required CPW discontinuity, two different groups of impressed current source, denoted as $I_{1a}, I_{2a}; I_{1b}, I_{2b}$, are

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applied. Then two resulting groups of port voltage, $V_{1a}, V_{2a}, V_{1b}, V_{2b}$ can be obtained by MoM. As a result, the admittance parameters of the equivalent circuit network can be formulated through the equivalent voltages and currents at the designated reference planes #1 and #2 as shown in Figure2.2. They can be deduced from equation (2.8) as equation (2.11):

$$\begin{cases} Y_{11} = (I'_{1a} V'_{2b} - I'_{1b} V'_{2a}) / (V'_{1a} V'_{2b} - V'_{1b} V'_{2a}) \\ Y_{12} = (I'_{2a} V'_{2b} - I'_{2b} V'_{2a}) / (V'_{1a} V'_{2b} - V'_{1b} V'_{2a}) \\ Y_{21} = (I'_{1a} V'_{1b} - I'_{1b} V'_{1a}) / (V'_{1b} V'_{2a} - V'_{1a} V'_{2b}) \\ Y_{22} = (I'_{2a} V'_{1b} - I'_{2b} V'_{1a}) / (V'_{1b} V'_{2a} - V'_{1a} V'_{2b}) \end{cases} \quad (2.11)$$

Figure 2.5 Susceptance derived by momentum simulation and MoM-SOC method

To justify this MoM-SOC advantage over commercial planar EM simulator, open series gap discontinuity shown in Figure 2.1 is taken as an example, the equivalent circuit susceptance calculated by one of EM software simulators (momentum in ADS) and MoM-SOC method, at two reference planes P_1 and P_2 , are presented in Figure 2.5. It

can be found that the result is obviously wrong without using SOC technique. This strongly proves the necessity of de-embedded technique.

2.4 Conclusion

In this chapter, the method of moments (MoM) algorithm and short-open calibration (SOC) techniques are described with regarding to two-port CPW discontinuity. Two CPW “short” and “open” elements are formulated in the MoM to accurately simulate the ideal open- and short-end blocks according to the transmission line theory. Thus, two-port circuit parameters can be extracted in terms of admittance matrix. This methodology will be used in next chapter to obtain different CPW discontinuities for MMIC application.

Chapter 3

Modeling of CPW Discontinuities

Abstract: In this chapter, a magnetic field integral equation (MFIE) is solved by executing method of moment in spectral domain. According to the MoM-SOC technique in chapter 2, circuit parameters of the core CPW discontinuity are then extracted. Finally, some typical CPW discontinuities to be used in the next chapters are analyzed through my own developed program.

3.1 Introduction

In recent decades, much more research interests have been aroused to coplanar waveguide (CPW) integrated circuit design due to its many advantages over conventional microstrip line [1]. A good understanding of CPW discontinuities' circuit behavior is essential for design success. The typical discontinuities include short end, open end, series gap, series or shunt stub, interdigital capacitive or inductive coupling element.

Chapter3: Modeling of CPW Discontinuities

Over the last 20 years, many theoretical and experimental modeling techniques were carried out [7, 11, 28, 30, 34, 37, 40, 45, 81-84]. Both quasi-static as well as fullwave approaches have been used to set up a library for computer-aided design (CAD) simulation tools of different CPW structures [75, 86]. Also, some commercial softwares based on fullwave method of moments [48, 50] are also available in recent years for circuit block simulation. However, the design information suitable for CAD tools is still lacking for most commonly used discontinuities. The reason seems to be that there are a large number of parameters in CPW discontinuities to be determined, therefore, it is very difficult to develop expressions covering all the useful design information. To my knowledge, only a few closed-form formulae have been published for most simple open and short circuit discontinuities under very strict conditions[17, 87, 88].

In this chapter, six typical CPW discontinuities circuit performances will be analyzed via their equivalent π - or T- type circuit network by my own developed MoM-SOC software. A number of accurate frequency-related curves with varied geometry parameters are exhibited to give a comprehensive guideline for further circuit block design.

3.2 Magnetic Field Integral Equation (MEIE)

Figure 3.1(a) shows the cross-section of CPW structure formed on the multilayered dielectric substrates. Figure 3.1(b) illustrates its alternative equivalent structure in

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which tangential electric field on the aperture (or slot) is perceived as a pair of positive and negative magnetic current density, i.e. $+\bar{M}$ and $-\bar{M}$ above and below the perfect metallic plate respectively.

Figure 3.1 (a) cross-section of CPW structure (b) its equivalent structure

Scattering magnetic field in the up- and down-region of the aperture interface ($z=0$) can be expressed in terms of the above two magnetic currents through the dyadic Green's functions $\bar{\bar{G}}_{HM}^+(\bar{r}, \bar{r}_s)$ and $\bar{\bar{G}}_{HM}^-(\bar{r}, \bar{r}_s)$

$$\bar{H}^{scat}(\bar{r}) = \begin{cases} \iint_{aperture} \bar{\bar{G}}_{HM}^+(\bar{r}, \bar{r}_s) \cdot (+\bar{M}(\bar{r}_s)) ds & (z > 0) \\ \iint_{aperture} \bar{\bar{G}}_{HM}^-(\bar{r}, \bar{r}_s) \cdot (-\bar{M}(\bar{r}_s)) ds & (z < 0) \end{cases} \quad (3.1)$$

By enforcing the boundary conditions over the aperture surface in which tangential magnetic fields must be continuous, the total magnetic field above the aperture must be equal to that below the aperture, such that,

$$\hat{z} \times (\bar{H}^{inc}(\bar{r}^+) + \bar{H}^{scat}(\bar{r}^+)) = \hat{z} \times (\bar{H}^{inc}(\bar{r}^-) + \bar{H}^{scat}(\bar{r}^-)) \quad (3.2)$$

$$\bar{J}^{inc}(\bar{r}) + \hat{z} \times (\bar{H}^{scat}(\bar{r}^+) - \bar{H}^{scat}(\bar{r}^-)) = 0 \quad (3.3)$$

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in which $\vec{H}^{inc}(\vec{r}^+)$ and $\vec{H}^{inc}(\vec{r}^-)$ indicate the incident or impressed fields, and they can be put together to formulate the equivalent electric current density source.

$$\vec{J}^{inc}(\vec{r}) = \hat{z} \times (\vec{H}^{inc}(\vec{r}^+) - \vec{H}^{inc}(\vec{r}^-)) = 0$$

substituting equation (3.1) to (3.3), we can have

$$\vec{J}^{inc}(\vec{r}) + \hat{z} \times \iint_{aperture} \overline{\overline{G}}_{HM}(\vec{r}, \vec{r}_s) \cdot (\vec{M}(\vec{r}_s)) dS = 0 \quad (3.4)$$

where

$$\overline{\overline{G}}_{HM}(\vec{r}, \vec{r}_s) = \overline{\overline{G}}_{HM}^+(\vec{r}, \vec{r}_s) + \overline{\overline{G}}_{HM}^-(\vec{r}, \vec{r}_s)$$

For the two-port CPW circuits, the integral region of this MFIE can be further classified into the two separate parts: two feeding lines and CPW circuit under modeling.

$$\sum_{j=1}^2 [\vec{J}^{inc}(\vec{r}) + \hat{z} \times \iint_{line(j)} \overline{\overline{G}}_{HM}(\vec{r}, \vec{r}_s) \cdot (\vec{M}_j(\vec{r}_s)) dS] + \hat{z} \times \iint_{circuit} \overline{\overline{G}}_{HM}(\vec{r}, \vec{r}_s) \cdot (\vec{M}(\vec{r}_s)) dS = 0 \quad (3.5)$$

3.3 Dyadic Spectral Green's Functions

One dimensionally inhomogeneous multilayer planar structure (dielectric or magnetic media) plays an important role in microwave and millimeter-wave planar transmission lines and antennas. One of several useful methods for the analysis of such structures is the full-wave (or full-electromagnetic) spectral domain technique, which has

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demonstrated excellent performance in electromagnetic simulation in both microwave and millimeter-wave bands. In this technique, integral equations for unknown electric and magnetic current distributions on planar lines and antennas are first obtained by using dyadic spectral Green's functions. Hence, the dyadic spectral Green's functions are a determinant factor in considering whether the full-wave spectral-domain technique can be used to solve a practical problem of planar circuits and antennas.

The spectral Green's functions for planar structures have been studied by a number of authors. They can be divided into three groups. One is called the vector potential method. Extending this method to problems of more than two-layered structures becomes too complicated and an iterative algorithm is proposed to take additional layers into consideration. The second method is known as the spectral domain impedance method, which is a spectral domain extension of the classical works on transverse field equations. Hence, the method is limited to two-dimensional problems. The third method is called the scattering superposition method. This procedure may become cumbersome because of the complex vector nature of electromagnetic waves.

Now I use the method that is presented for the complete determination of the dyadic spectral Green's functions directly from Maxwell's equations in the spectral domain [89]. Dyadic Green's functions represent the induced electric and/or magnetic (dyadic) fields of delta electric or magnetic (dyadic) current source in a multilayered structure, which is shown in Figure 3.2.

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Figure 3.2 One dimensionally inhomogeneous multilayer structures excited by three dimensional electrical or magnetic point sources

According to the Maxwell's equations:

$$\nabla \times \vec{E}(\vec{r}) = -j\omega\mu_i \vec{H}(\vec{r}) - \vec{M}(\vec{r}) \quad (3.6a)$$

$$\nabla \times \vec{H}(\vec{r}) = j\omega\varepsilon_i \vec{E}(\vec{r}) + \vec{J}(\vec{r}) \quad (3.6b)$$

and the equations

$$\nabla \cdot (\nabla \times \vec{H}(\vec{r})) = 0$$

$$\nabla \cdot (\nabla \times \vec{E}(\vec{r})) = 0$$

we can obtain

$$\nabla \cdot [j\omega\varepsilon_i \vec{E}(\vec{r})] = -\nabla \cdot \vec{J}(\vec{r}) \quad (3.6c)$$

$$\nabla \cdot [j\omega\mu_i \vec{H}(\vec{r})] = -\nabla \cdot \vec{M}(\vec{r}) \quad (3.6d)$$

Using the standard vector analysis, the electric and magnetic fields can be expressed in terms of the current sources as

$$\vec{E}(\vec{r}) = \int_v \vec{\overline{G}}_{EJ}(\vec{r}, \vec{r}_s) \cdot \vec{J}(\vec{r}_s) d\vec{r}_s + \int_v \vec{\overline{G}}_{EM}(\vec{r}, \vec{r}_s) \cdot \vec{M}(\vec{r}_s) d\vec{r}_s \quad (3.7a)$$

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$$\vec{H}(\vec{r}) = \int_v \vec{\bar{G}}_{HM}(\vec{r}, \vec{r}_s) \cdot \vec{M}(\vec{r}_s) d\vec{r}_s + \int_v \vec{\bar{G}}_{HJ}(\vec{r}, \vec{r}_s) \cdot \vec{J}(\vec{r}_s) d\vec{r}_s \quad (3.7b)$$

where the dyadic Green's functions satisfy the dyadic version of Maxwell's equations, the magnetic current source dyadic Green's functions are expressed as below for modeling the CPW structures with unknown equivalent magnetic current density over the aperture or slot interface:

$$\nabla \times \vec{\bar{G}}_{EM}(\vec{r}, \vec{r}_s) = -j\omega\mu_i \vec{\bar{G}}_{HM}(\vec{r}, \vec{r}_s) - \vec{\bar{I}}\delta(\vec{r} - \vec{r}_s) \quad (3.8a)$$

$$\nabla \times \vec{\bar{G}}_{HM}(\vec{r}, \vec{r}_s) = j\omega\epsilon_i \vec{\bar{G}}_{EM}(\vec{r}, \vec{r}_s) \quad (3.8b)$$

$$\nabla \cdot [j\omega\epsilon_i \vec{\bar{G}}_{EM}(\vec{r}, \vec{r}_s)] = 0 \quad (3.8c)$$

$$\nabla \cdot [j\omega\mu_i \vec{\bar{G}}_{HM}(\vec{r}, \vec{r}_s)] = -\nabla \cdot \vec{\bar{I}}\delta(\vec{r} - \vec{r}_s) \quad (3.8d)$$

where $\vec{\bar{I}}$ is a unit dyadic.

The transverse Fourier transformation is defined as:

$$\vec{\bar{G}}(\vec{r}, \vec{r}_s) = \frac{1}{4\pi^2} \vec{\bar{G}}(z, z_s; \alpha, \beta) e^{-j\alpha(x-x_s) - j\beta(y-y_s)} d\alpha d\beta \quad (3.9a)$$

and

$$\vec{\nabla} = -jk_t + \frac{d}{dz} \hat{z} \quad (3.9b)$$

$$\vec{k}_t = \alpha\hat{x} + \beta\hat{y}$$

$$\vec{k}_t = k_t \hat{k}_t = \sqrt{\alpha^2 + \beta^2} \left(\frac{\alpha}{\sqrt{\alpha^2 + \beta^2}} \hat{x} + \frac{\beta}{\sqrt{\alpha^2 + \beta^2}} \hat{y} \right) \quad (3.9c)$$

The dyadic Green's functions of magnetic current sources in spectral domain are expressed as:

$$\vec{\nabla} \times \vec{\bar{G}}_{EM}(\vec{k}_t, z, z_s) = -j\omega\mu_i \vec{\bar{G}}_{HM}(\vec{k}_t, z, z_s) - \vec{\bar{I}}\delta(z - z_s) \quad (3.10a)$$

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$$\tilde{\nabla} \times \tilde{\tilde{G}}_{HM}(\vec{k}_t, z, z_s) = j\omega\epsilon_i \tilde{\tilde{G}}_{EM}(\vec{k}_t, z, z_s) \quad (3.10b)$$

$$\tilde{\nabla} \cdot [j\omega\epsilon_i \tilde{\tilde{G}}_{EM}(\vec{k}_t, z, z_s)] = 0 \quad (3.10c)$$

$$\tilde{\nabla} \cdot [j\omega\mu_i \tilde{\tilde{G}}_{HM}(\vec{k}_t, z, z_s)] = -\tilde{\nabla} \cdot \bar{I} \delta(z - z_s) \quad (3.10d)$$

The vector spectral eigenfunctions are defined, for TE modes, as

$$\tilde{M}^h(\vec{k}_t, z) = \frac{-1}{jk_t} \tilde{\nabla} \times \hat{z} V^h(z) = V^h(z) (\hat{k}_t \times \hat{z}) \quad (3.11a)$$

$$\tilde{N}^h(\vec{k}_t, z) = \frac{-1}{j\omega\mu_i} \tilde{\nabla} \times \tilde{M}^h(\vec{k}_t, z) = I^h(z) \hat{k}_t - \frac{k_t}{\omega\mu_i} V^h(z) \hat{z} \quad (3.11b)$$

for TM modes:

$$\tilde{M}^e(\vec{k}_t, z) = -I^e(z) (\hat{k}_t \times \hat{z}) \quad (3.12a)$$

$$\tilde{N}^e(\vec{k}_t, z) = V^e(z) \hat{k}_t - \frac{k_t}{\omega\epsilon_i} I^e(z) \hat{z} \quad (3.12b)$$

where $\tilde{M}^h$ and $\tilde{N}^e$ are the electric eigen-fields of the modes, $\tilde{M}^e$ and $\tilde{N}^h$ are the magnetic eigenfunctions of the modes. By substituting equation (3.11) and (3.12) to a source free region homogeneous Maxwell's equations it is easy to verify that V^e, I^e, V^h, I^h satisfied the homogeneous transmission-line equations:

$$\begin{aligned} \frac{d}{dz} V(z) + jk_z(z) Z(z) I(z) &= 0 \\ \frac{d}{dz} I(z) + jk_z(z) Y(z) V(z) &= 0 \end{aligned} \quad (3.13)$$

and

$$Z(z) = \frac{1}{Y(z)} = \begin{cases} Z^e = \frac{k_z(z)}{\omega\epsilon_i(z)} \\ Z^h = \frac{\omega\mu_i(z)}{k_z(z)} \end{cases} \quad k_z^2(z) = \omega^2 \mu_i \epsilon_i - k_t^2$$

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Dyadic spectral domain eigenfunctions for TE and TM modes

$$\begin{aligned}
 (\tilde{\tilde{M}}\tilde{\tilde{M}})^{e,h}(\vec{k}_t, z, z_s) &= \tilde{\tilde{M}}^{e,h}(\vec{k}_t, z) \tilde{\tilde{M}}^{e,h}(-\vec{k}_t, z_s) \\
 (\tilde{\tilde{M}}\tilde{\tilde{N}})^{e,h}(\vec{k}_t, z, z_s) &= \tilde{\tilde{M}}^{e,h}(\vec{k}_t, z) \tilde{\tilde{N}}^{e,h}(-\vec{k}_t, z_s) \\
 (\tilde{\tilde{N}}\tilde{\tilde{M}})^{e,h}(\vec{k}_t, z, z_s) &= \tilde{\tilde{N}}^{e,h}(\vec{k}_t, z) \tilde{\tilde{M}}^{e,h}(-\vec{k}_t, z_s) \\
 (\tilde{\tilde{N}}\tilde{\tilde{N}})^{e,h}(\vec{k}_t, z, z_s) &= \tilde{\tilde{N}}^{e,h}(\vec{k}_t, z) \tilde{\tilde{N}}^{e,h}(-\vec{k}_t, z_s)
 \end{aligned} \tag{3.14}$$

the modal amplitudes with variables z and z_s are combined and re-defined as

$$\begin{aligned}
 W^{e,h}(z, z_s) &= V^{e,h}(z) V^{e,h}(z_s) \\
 I_V^{e,h}(z, z_s) &= I^{e,h}(z) V^{e,h}(z_s) \\
 V_I^{e,h}(z, z_s) &= V^{e,h}(z) I^{e,h}(z_s) \\
 II^{e,h}(z, z_s) &= I^{e,h}(z) I^{e,h}(z_s)
 \end{aligned} \tag{3.15}$$

then the magnetic Green's functions in spectral-domain can be expressed. Since $\tilde{\tilde{G}}_{EM}$ is solenoidal, $\tilde{\tilde{G}}_{EM}$ can be expanded completely in terms of the dyadic spectral eigenfunctions $(\tilde{\tilde{M}}\tilde{\tilde{N}})^h, (\tilde{\tilde{N}}\tilde{\tilde{M}})^e$.

$$\begin{aligned}
 \tilde{\tilde{G}}_{EM}(\vec{k}_t, z, z_s) &= (\tilde{\tilde{M}}\tilde{\tilde{N}})^h(\vec{k}_t, z, z_s) + (\tilde{\tilde{N}}\tilde{\tilde{M}})^e(\vec{k}_t, z, z_s) \\
 &= -V_I^h(z, z_s)(\hat{k}_t \times \hat{z})\hat{k}_t - \frac{k_t}{\omega\mu_i} W^h(z, z_s)(\hat{k}_t \times \hat{z})\hat{z} + V_I^e(z, z_s)\hat{k}_t(\hat{k}_t \times \hat{z}) \\
 &\quad - \frac{k_t}{\omega\epsilon_i} II^h(z, z_s)\hat{z}(\hat{k}_t \times \hat{z})
 \end{aligned}$$

$$\begin{aligned}
 \tilde{\tilde{G}}_{HM}(\vec{k}_t, z, z_s) &= -\frac{1}{j\omega\mu_i} [\tilde{\nabla} \times \tilde{\tilde{G}}_{EM}(\vec{k}_t, z, z_s) + \bar{I}\delta(z - z_s)] \\
 &= -II^e(z, z_s)(\hat{k}_t \times \hat{z})(\hat{k}_t \times \hat{z}) - II^h(z, z_s)\hat{k}_t\hat{k}_t \\
 &\quad + \frac{k_t}{\omega\mu_i(z)} V_I^h(z, z_s)\hat{z}\hat{k}_t - \frac{k_t}{\omega\mu_i(z_s)} I_V^h(z, z_s)\hat{k}_t\hat{z} \\
 &\quad + \frac{k_t^2}{\omega^2\mu_i(z)\mu_i(z_s)} W^h(z, z_s)\hat{z}\hat{z} - \frac{\delta(z - z_s)}{j\omega\mu_i(z)} \hat{z}\hat{z}
 \end{aligned}$$

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where the modal amplitudes satisfy two sets of inhomogeneous transmission-line equations:

Transverse magnetic current:

$$\begin{aligned} \frac{d}{dz} V_I^{e,h}(z, z_s) + jk_z(z) Z^{e,h}(z) I I^{e,h}(z, z_s) &= \delta(z - z_s) \\ \frac{d}{dz} I I^{e,h}(z, z_s) + jk_z(z) Y^{e,h}(z) V_I^{e,h}(z, z_s) &= 0 \end{aligned} \quad (3.16)$$

Vertical magnetic current:

$$\begin{aligned} \frac{d}{dz} W^h(z, z_s) + jk_z(z) Z^h(z) I_V^h(z, z_s) &= 0 \\ \frac{d}{dz} I I^{e,h}(z, z_s) + jk_z(z) Y(z) V_I^h(z, z_s) &= \delta(z - z_s) \end{aligned} \quad (3.17)$$

where $I I^{e,h} = I$; $V_I^{e,h} = V$; $W^h = V$; $I_V^h = I$;

Figure 3.3 Equivalent TE and TM circuit's models for one-dimensionally inhomogeneous multilayer planar structures excited by a three-dimensional magnetic point source. (a) TE and TM transverse magnetic point source excitation. (b) TE modes for vertical magnetic point source excitation.

3.4 Dyadic Green's Function for a Single-layered CPW

Structure

In my research, the dyadic Green's function of a magnetic current source needs to be deduced in conjunction with the single-layered CPW structure. As described above, a pair of magnetic current density is put above and below the metallic plate instead of electric field over the aperture surface as shown in Figure 3.4 (a). Its equivalent circuit is described in Figure 3.4 (b), in which $+\vec{M}$ and $-\vec{M}$ with the delta-function is expressed as a pair of the unit voltage sources, i.e. $+1V$ and $-1V$.

Figure 3.4 (a) a single-layered CPW structure (b) its equivalent circuit

In this case, the two dyadic eigenfunctions, II^e and II^h can be expressed as electric currents. On the aperture interface, they can be derived in terms of linear superposition of the up-and down-currents I^{up} and I^{down}

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$$\begin{aligned}
 II^{e,h} &= I = I^{up} + I^{down} = \frac{1}{z^{up}} + \frac{1}{z^{down}} \\
 z^{up} &= z_0^{e,h} \\
 z^{down} &= z_{in} = z_1^{e,h} \cdot \frac{z_0^{e,h} \cosh jk_z h + z_1^{e,h} \sinh jk_z h}{z_1^{e,h} \cosh jk_z h + z_0^{e,h} \sinh jk_z h} \\
 II^{e,h} &= \frac{1}{z_0^{e,h}} + \frac{z_1^{e,h} \cosh jk_z h + z_0^{e,h} \sinh jk_z h}{z_0^{e,h} \cosh jk_z h + z_1^{e,h} \sinh jk_z h} \cdot \frac{1}{z_1^{e,h}} \\
 \widetilde{\widetilde{G}}_{HM}(\bar{k}_t, z, z_s) &= -II^e(z, z_s)(\hat{k}_t \times \hat{z})(\hat{k}_t \times \hat{z}) - II^h(z, z_s)\hat{k}_t \hat{k}_t
 \end{aligned} \tag{3.18}$$

By substituting equation (3.9c) into equation (3.18), we can obtain

$$\begin{aligned}
 \widetilde{\widetilde{G}}_{HM}(\bar{k}_t, z, z_s) &= G_{xx} \hat{x} \hat{x} + G_{xy} \hat{x} \hat{y} + G_{yx} \hat{y} \hat{x} + G_{yy} \hat{y} \hat{y} \\
 G_{xx} &= \frac{-(II^e \beta^2 + II^h \alpha^2)}{\alpha^2 + \beta^2}; & G_{xy} &= \frac{\alpha \beta (II^e - II^h)}{\alpha^2 + \beta^2} \\
 G_{yx} &= \frac{\alpha \beta (II^e - II^h)}{\alpha^2 + \beta^2}; & G_{yy} &= \frac{-(II^e \alpha^2 + II^h \beta^2)}{\alpha^2 + \beta^2}
 \end{aligned}$$

3.5 Frequency Dispersive Characteristic Impedance of CPW Uniform Line

As the most basic parameters for MMIC design, the frequency-related characteristics, i.e., characteristic impedance and effective dielectric constant of uniform CPW line definition is carried out in terms of the transverse field quantities from a two-dimensional (2-D) numerical calculation under the assumption of the conductor thickness t is zero. In [82], closed form formulae are given based on the conformal

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mapping method. As detailed in [82], this approach assumes that all the dielectric interferences including slots, can be replaced by magnetic walls, thus, half-plane above and below the metallization plane of consider structure (as shown in Figure 1.1) can be analyzed separately for line capacitance. The total line capacitance is then the sum of the two capacitances.

So, effective dielectric constant ϵ_{reff} and characteristic impedance Z_C can be expressed in terms of elliptic integrate k_1 :

$$\epsilon_{\text{reff}} = 1 + \frac{\epsilon_r}{2} \cdot \frac{K(k_2)}{K'(k_2)} \cdot \frac{K'(k_1)}{K(k_1)} \quad (3.19a)$$

$$Z_C = \frac{30\pi}{\sqrt{\epsilon_{\text{reff}}}} \cdot \frac{K'(k_1)}{K(k_1)} \quad (3.19b)$$

where, the argument of the elliptic integrals is obtained as:

$$k_1 = a/b = S/(S + 2W)$$

$$k_2 = \frac{\sinh(\pi a / 2h)}{\sinh(\pi b / 2h)}$$

one of accurate and simple expression of ratio K/K' can be obtained in [81]:

$$\frac{K(k)}{K'(k)} = \frac{\pi}{\ln[2(1 + \sqrt{k'})/(1 - \sqrt{k'})]} \quad \text{For } 0 \leq k \leq 0.707$$

$$\frac{K(k)}{K'(k)} = \frac{1}{\pi} \ln[2(1 + \sqrt{k})/(1 - \sqrt{k})] \quad \text{For } 0.707 \leq k \leq 1$$

$$k' = \sqrt{1 - k^2} .$$

But, actually, the impedance exhibits frequency dispersive characteristics. It was

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originated in [80], and has been extensively investigated [90-92] via three-dimensional (3-D) method of moments (MoM) together with numerical de-embedding techniques.

Next, they will be defined and characterized with resorting to the network equivalence as a finitely extended CPW line in three-dimension (3D) MoM-SOC technique introduced in Chapter 2. Firstly, three cascaded CPW line sections are arranged for unified 3-D definition of CPW characteristic impedances. The middle CPW line section with length L_p is considered as the core DUT as described in [93] while sections on either side are considered as feeding lines. As a result, CPW transmission line network parameters can be efficiently extracted and expressed as two-port ABCD matrix elements:

$$Z_c = \sqrt{\frac{B}{C}}$$

$$\varepsilon_{\text{reff}} = \left[\frac{c}{\omega L_p} \left(n\pi + a \tan\left(\frac{B}{jAZ_c}\right) \right) \right]^2 \quad (3.20)$$

In which, c is the light velocity in free space, n is the integer to be determined.

Figure 3.5 depicts the calculated characteristic impedance of a uniform CPW line on substrate thickness 25mil and dielectric constant 10.8 with fixed length $L_p=250\text{mil}$ under different dimensions of slot width s and strip width w . All the results are very closed to conformal mapping in equation (3.19a) noted with circle at lower frequency (below 8GHz). In fact, they appear to gradually rise up with frequency, thus exhibiting their frequency related behaviors. Further we can find that characteristic impedance increases with reduced strip width w , as shown in Figure 3.5 (a) or extended slot width

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s in Figure 3.5 (b).

(a)

(b)

Figure 3.5 Characteristic impedance versus frequency with different (a) strip width w (b) slot width s

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Figure 3.6 Effective dielectric constant versus frequency with different (a) strip width w (b) slot width s

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Figure 3.6 (a) and Figure 3.6 (b) describe the calculated effective dielectric constant of a uniform CPW line under different w and s within the frequency range 5GHz to 15GHz, respectively. They all increase smoothly with frequency. As widening s or w , such effective dielectric constants are observed to decrease by fixing other parameters.

3.6 Some Typical CPW Discontinuities

In this section, some CPW discontinuities used in this thesis will be analyzed in terms of T- or π - reactance or susceptance network by using my developed MoM-SOC software.

3.6.1 Capacitive Coupling Element

A. Conventional Lumped Open End Coupling Element

Figure 3.7 (a) depicts the physical layout of a conventional open-end capacitive coupling element connected by two uniform feed lines. The discontinuity terminated by reference planes P_1 and P_2 are modeled as π -type network with two equal shunt susceptance Y_p at either side and one series susceptance Y_g in the middle. Such type of discontinuity has been theoretically investigated using the full-wave method of moments [96] and finite difference method [40], respectively. Here, firstly, the MoM-SOC technique is used to extract the admittance parameters of this structure again to

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ensure the accuracy.

(a)

(b)

Figure 3.7 Geometry of open-end capacitive coupling element and its equivalent π -type network topology. (a) layout (b) equivalent circuit model

The transform of admittance parameters to equivalent π type admittance ($Y = G + j\omega C$) network can be derived using equation (3.21)

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} Y'_{11} & Y'_{12} \\ Y'_{21} & Y'_{22} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} \quad (3.21)$$

thus the equivalent series and shunt admittance are available :

$$\begin{aligned} Y_p &= Y'_{11} + Y'_{12} = G_p + j\omega C_p \\ Y_g &= -Y'_{12} = G_g + j\omega C_g \end{aligned} \quad (3.22)$$

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In equation (3.22), G_p and G_g represent the radiation loss of corresponding susceptance. Shunt admittance ($Y_p = G_p + jB_p$) and series admittance ($Y_g = G_g + jB_g$) as a function of frequency, with different gap width g as parameters, are shown in Figure 3.8(a) and 3.8(b) respectively. In this section, all the CPW elements are formed on the material with dielectric permittivity of 10.8, substrate thickness of 25mil, CPW feed line keeps as $50\ \Omega$ with central strip conductor width w of 50mil, and slot width s of 15mil. Shunt conductance (G_p) and series conductance (G_g) are found negligibly small compared to susceptance at our concerned frequency range. It indicates that the radiation loss is very small. And also, series and shunt susceptance behave quasi-linearly in the frequency range (5-10GHz). Thus, generally, it can be modeled as one series capacitance C_g and two equal shunt capacitances C_p , they are found to be virtually frequency independent in the considered frequency range. As widening the gap width g , the shunt susceptance B_p is observed to increase while the series susceptance B_g decreases. This somewhat illustrates that the coupling effect of two-port gap structure becomes stronger if decreasing the gap width at fixed other parameters. Thus, different coupling degree can be obtained by simply adjusting the gap width g . Moreover, the shunt or series capacitance C_p/C_g can be adjusted by slot width or central conductor width. Figure 3.9 denotes extracted shunt or series capacitance versus slot width s under the fixed transverse distance ($w+2s=110\text{mil}$) at 5GHz. It can be observed that the shunt capacitance C_p is much smaller than that of series capacitance C_g that decreases significantly with stretched s .

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Figure 3.8 Frequency dispersive admittance as a function of gap distance g (a) shunt admittance $Y_p = G_p + jB_p$ (b) series admittance $Y_g = G_g + jB_g$

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Figure 3.9 Extracted lumped-network parameters of shunt-C and series-C with fixed transversal width $w+2s$

B. Enhanced Distributed Capacitive Coupling Element

Figure 3.10 Geometry of enhanced distributed capacitive coupling element

Figure 3.10 shows the geometry of enhanced distributed capacitive coupling element, which is constituted by extending the conventional capacitive element stub. But the

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frequency dispersive responses are quite different. An equivalent π -type network is also used to analyze this structure. Many parameters will affect the shunt and series susceptance such as the gap width g , slot width l_1 , stub length l_3 and strip gap l_2 .

Figure 3.11 Frequency dispersive response of shunt and series susceptances (B_p and B_g) as a function of gap width g

Firstly, by choosing quite small size of l_1 and l_3 , the series susceptance B_g and shunt susceptance B_p versus frequency as a function of g are plotted in Figure 3.11. Three curves of B_p linearly increase with frequency, thus generally modeled as frequency-independent shunt capacitance C_p in the denoted frequency range. Meanwhile, the series susceptance B_g increases more rapidly with frequency. Similar to lumped capacitive coupling element, as enlarging the gap width g , the shunt susceptance B_p increases while the series susceptance B_g decreases.

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Figure 3.12 Frequency dispersive response of (a) shunt admittance Y_p and (b) series admittance Y_g as a function of stub length l_3

Special attention will be paid to this structure with their usually significant coupling

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strength as well as series resonant behavior. This phenomenon is obviously with the stub length increasing. Actually, the series susceptance B_g of π network can be modeled as a series LCR resonator. To illustrate the resonator characteristics, the complete admittances $Y_g (=G_g + jB_g)$ are plotted under different three length of l_3 by fixed l_1 of 15mil, l_2 of 10mil and g of 5mil over the frequency range (2-10GHz) as shown in Figure 3.12 (b). It indicates that the series susceptance (B_g) behaves quasi-linearly in the lower frequency range and then an abrupt increment occurs as frequency increases and then fall down quickly to negative value and then goes up by passing zero around resonator point. The series conductance (G_g) is extremely small at lower frequency and then it goes up rapidly. As frequency increases further, B_g drops quickly while G_g is raised to a certain maximum value, thus suggesting the RLC resonator phenomenon. If the stub length widen, the peak point is shifted up toward lower frequency, thereby showing a great coupling enhancement. As shown in Figure 3.12 (a), the shunt susceptance behaviors linear function as frequency regardless of different length l_3 , and the shunt conductance are negligibly small. Thus constructing two equal frequency-independent C_p are constructed at either side of π type network.

Other parameters also have an effect on the coupling strength and resonant frequency of this structure. Figure 3.13 shows the B_p and B_g under different length l_2 by fixing $l_1=15\text{mil}$, $l_3=20\text{mil}$, $G=5\text{mil}$. As the strip gap l_2 is widened from 10mil to 30mil, both susceptance increases, the series susceptance resonant frequency is observed to move down to lower value. Figure 3.14 depicts the B_p and B_g under different length l_1 by choosing unchanged l_2 as 10mil, l_3 as 20mil and g as 5mil. As l_1 is widened, both of

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them increases and the peak of B_g moves toward lower frequency.

Figure 3.13 Dispersive susceptance as a function of l_2

Figure 3.14 Dispersive susceptance as a function of l_1

C. Open-end CPW Series Stub

(a)

(b)

Figure 3.15 Geometry and its equivalent circuit network for open end coplanar waveguide series stub (a) physic layout (b) equivalent asymmetric π type network

An open circuited stub constituted by an interdigital finger structure of central strip conductor of CPW is shown in Figure 3.15 (a). Asymmetric π type circuit network, with one series susceptance Y_c in the center, shunt susceptance Y_a and Y_b in the left and right, respectively, is used to analyze the circuit characteristics at two reference planes P_1 and P_2 . In fact, this structure can be accurately modeled as a series capacitance at very small length d_3 [42] (less than $\lambda_g/10$), but this is not valid for a general case [94] where this length is stretched to realize the large coupling degree for wideband bandpass filter. In this case, the series admittance Y_c is actually formed with LC

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resonator as demonstrated in [94]. Next, let us see the frequency dispersive circuit performance by choosing different parameters.

Figure 3.16 Extracted curves of equivalent π network based frequency dispersive susceptance as a function of mean stub length d_3

Figure 3.16 plots a number of curves for susceptance as a function of frequency under the different stub length d_3 . Within the interested frequency range (2-5GHz), the shunt susceptance B_a/ω and B_b/ω almost keep constant at lower frequency below 3GHz, while with frequency increasing, B_b/ω shows slightly decreases after 4GHz, while B_a/ω increases greatly than that in lower frequency range. The series susceptance B_c exhibits gradually increase with frequency. Further, we can find at certain frequency, left shunt equivalent lumped capacitance B_a/ω and central series equivalent capacitance B_c/ω gradually increases with extended stub length d_3 , while

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right shunt equivalent capacitance B_b / ω exhibits slight decrease.

Figure 3.17 Frequency dispersive susceptance as a function of coupled distance w_2

The circuits' parameters as the function of slot width s and coupled slot width w_2 can be used for wideband bandpass filter design. Figure 3.17 depicts the frequency-dependant curves under three different w_2 at fixed dimensions: central strip width=50mil, $w_1=5$ mil, stub length $d_3=80$ mil and slot width $s=15$ mil. It can be observed that shunt equivalent capacitance B_a / ω and B_b / ω are almost constant during considered frequency, while the series equivalent capacitance B_c / ω increases gradually with frequency. Moreover, series capacitance exhibits much more great decrease than that of right shunt capacitance with extended coupled distance w_2 , thus showing a larger series capacitance with smaller parasitic capacitance with small coupled space.

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Figure 3.18 Frequency dispersive susceptance as a function of slot width s

Figure 3.18 plots the susceptance curves under three different slot width s . With widened s , two shunt equivalent capacitances exhibit decreasing value, while series capacitance increases, thus showing a great enhancement of coupling behavior as well as reduction of parasitic effects caused by shunt capacitances.

3.6.2 Inductive Coupling Element

A. Conventional End Coupled Element

Figure 3.19(a) shows the geometry of CPW end coupled inductive coupling element formed by grounding two port slots with distance G . Thus, its equivalent circuit appears as a T network at reference place P_1 and P_2 , as shown in Figure 3.19 (b), which consists

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of two series impedances ($Z_p = R_p + j\omega L_p$) and one shunt impedance ($Z_g = R_{12} + j\omega L_g$). An approximate method to estimate the shunt inductance has been investigated in the ribbon inductors [17]. However, the accuracy of this formula is quite poor, with errors on the order of 10 percent or greater for using the relatively wide width ribbons. An alternative approach to obtain the effective inductance of the structure is to conduct a series of experiments to measure the loaded quality factor and the transmission coefficient of a single section of resonator. It is impossible to be used to real design due to complex procedures. A full-wave method needs to be proposed to accurately extract the series or shunt inductance in terms of equivalent T type network.

Figure 3.19 Geometry and equivalent T-type network topology (a) layout (b) equivalent circuit model

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(a)

(b)

Figure 3.20 Frequency dispersive shunt and series impedance as a function of gap width G (a) series impedance Z_g (b) shunt impedance Z_p

The admittance matrix parameters of this structure can be obtained by the MoM-SOC

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method. The transform from impedance matrix of the CPW to the corresponding resistance and inductance in its equivalent circuit is expressed in equations (3.23):

$$\begin{aligned} Z_p &= Z_{11}' - Z_{12}' = R_p + j\omega L_p \\ Z_g &= Z_{21}' = Z_{12}' = R_g + j\omega L_g \end{aligned} \quad (3.23)$$

In this equation (3.23), R_p and R_g present the radiation-caused resistance, while L_p and L_g present series and shunt inductance. The extracted shunt and series impedance by our MoM-SOC as the function of frequency under different three gap width G are shown in Figure 3.20 (a) and Figure 3.20(b), respectively. Other dimensions are fixed at slot width $s=15\text{mil}$, central conductor width 50mil . Stable results are obtained with my program and the curve is smoothly changed to exhibit the good lumped element inductors.

Resistances show negligibly small compared with reactance. It can be observed that series resistance (R_p) is almost constant at lower frequency with fixed gap width (G) and increases gradually as frequency becomes larger, while the shunt resistance (R_g) increases almost linear with frequency. As shown in Figure 3.20, the shunt reactance X_g decreases with widened gap width G , while the series reactance X_p increases, thus demonstrating a gradually increasing inductive coupling with smaller gap width.

In addition, the slot width will take a great effect on circuit performances. Figure 3.21 depicts the shunt and series inductances as a function of slot width under fixed the total transversal length ($w+2s=110\text{mil}$) and gap width $G=5\text{mil}$. The series inductance L_p is found to increase slightly as s varies from 5mil to 45mil . The shunt inductance L_g is

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much larger than L_p and increases quickly with extended slot width s .

Figure 3.21 Equivalent lumped series and shunt inductance as a function of slot width s

B. Enhanced Distributed Inductive Coupling Element

Figure 3.22 Geometry of enhanced distributed inductive coupling element

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(a)

(b)

Figure 3.23 Frequency dispersive curves of T-type network parameters under different length d_1 (a) series impedance Z_p (b) shunt impedance Z_g

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Figure 3.22 shows the layout of enhanced distributed inductive coupling element formed by two short stubs with distance G . The coupling degree is certainly enhanced compared to conventional end-coupled structure by extending the strip conductor length (d_1 and d_2). A symmetrical T- type network with two equal series impedances Z_p and one shunt impedances Z_g still can be used as shown in Figure 3.19(b) to analyze this structure.

A number of numerical results with varied parameters will be plotted and discussed. The two CPW feed lines are 50Ω with 50mil strip width and 15mil slot width. Figure3.23 plots frequency dispersive shunt and series impedance under three different stub length d_1 , at the condition $G=5\text{mil}$, $d_2=10\text{mil}$, $d_3=15\text{mil}$. The resistance R_p and R_g are two orders less than that of the reactance, showing a negligible effect. The series reactance X_p and X_g exhibit linear response within denoted frequency range (2-10GHz). Both of them increase with extended stub length d_1 at fixed frequency, and the series reactance X_p is observed larger than shunt reactance X_g .

Figure 3.24 shows the frequency response under three different stub width d_3 , the series reactance X_p exhibits linear frequency response thus demonstrating equivalent inductor characteristics. The shunt reactance X_g linearly increases at lower frequency and accelerates at higher frequency. It also can be observed that enhancement of shunt reactance X_g is much smaller than that of series reactance X_p at certain frequency point as stub width d_3 widen. Figure 3.25 and 3.26 plots frequency related equivalent shunt and series inductance under different gap width G and stub length d_2 , respectively.

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Shunt inductance X_g and series inductance X_p increase with extended d_2 , while X_g reduces and X_p increases with G widened.

Figure 3.24 Frequency dispersive curves of T-type network parameters under different length d_3 (a) series impedance Z_p (b) shunt impedance Z_g

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Figure 3.25 Frequency dispersive curves of equivalent inductances as a function of gap width G

Figure 3.26 Frequency dispersive curves of equivalent shunt or series inductance as a function of d_2

C. Asymmetric Overlapped Inductive Coupling Element

Figure 3.27 Geometry and equivalent T-type circuit network topology of asymmetric overlapped coupling structure (a) layout (b) equivalent circuit network

An overlapped inductive coupling element composed of two coupled-slot pairs that are symmetric about the center of the CPW line is shown in Figure 3.27 (a). This topology was originally proposed by Menzel [97], that used symmetrical pairs of asymmetric coupled slots, for narrow-band filter design due to its advantages of being uniplanar, no requirement of air-bridges for ground equalization and suppressions of unwanted radiation. And later it was applied to the design of bandpass filters with 5-20% bandwidth on high-receptivity silicon [98]. In this part, an asymmetric T-type network is used to describe such structure with two series impedances Z_a and Z_b at two ends and one shunt impedance Z_c in the middle. A number of frequency response curves of these

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three impedances with varied coupled length T and coupled space d are given to thoroughly demonstrate the overlapped inductive element's circuit performances.

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Figure 3.28 Frequency dispersive equivalent T-type network curves under different coupled length T of (a) (b) self impedance (c) mutual impedance

Figure 3.28 plots the frequency dispersive curves of shunt and series impedance under different coupled length T. CPW feed line at left-hand side is 50Ω with 15mil slot width and 50mil strip width. The space d between coupled slots is 5mil. Resistances of both impedances are found less than 1 percent of corresponding reactance, thus indicating that the radiation of this structure is still negligible small. The self-reactance X_a and X_b exhibit linearly function of frequency to be equivalent as inductances with the considered frequency range and they increase as coupled length is enlarged. At smaller coupled length, the mutual reactance X_c is found to linearly increase with frequency so as to exhibit almost constant equivalent mutual inductance. As T further increases, such a mutual inductance seems to increase with faster velocity and its resistance becomes much larger than that of X_a and X_b .

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(a)

(b)

Figure 3.29 Frequency dispersive curves of reactance under different coupled width d

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Figure 3.29 plots frequency dispersive curves of reactance X_a , X_b and X_c under four different coupled space d , respectively. By reducing the coupling space, greater mutual inductance and less self inductance are obtained, thus showing a higher coupling effect. Based on the above two sets of numerical curves, the tight coupling can be achieved by either lengthening the coupled length T or reducing the coupled slot space d .

3.7 Conclusions

In this chapter, the magnetic field integral equation (MEIE) and self-calibrated MoM-SOC are described to analyze several CPW capacitive and inductive coupling discontinuities as well as uniform CPW transmission line characteristics and effective dielectric constant. Firstly, the characteristic impedance of uniform CPW line derived by MoM-SOC is compared with well-known conformal mapping method; the discrepancy is within 1% at lower frequency, which validates the accuracy of my developed program. Next, several two-port CPW capacitive or inductive coupling elements have been characterized by the MoM-SOC method in terms of equivalent π or T- type circuit networks, respectively. Comprehensive and accurate numerical results with varied different geometry parameters are given to provide a set of curves to design the complete CPW circuit blocks using the simple transmission line topology as will be implemented in the next chapters.

Chapter 4

Compact Quarter-Wavelength Bandpass Filters

Abstract: Accurate equivalent circuit models of several types of coplanar waveguide (CPW) coupling elements are presented and their parameters are derived by employing the short-open calibration (SOC) technique into the fullwave method of moments (MoM). Afterwards, the three sets of J- or K-inverter parameters versus frequency are extracted and obtained to demonstrate comparatively coupling degree of lumped capacitive and inductive gap-coupling as well as distributed overlapped-coupling structures. Next, these coupling elements are utilized together to design of compact quarter-wavelength CPW bandpass filters via simple cascaded network topology. Our optimized results are reasonably validated by ADS-simulation and experiment, both them exhibit good passband behavior with the broadened bandwidth and ultra-broad stopband.

4.1 Introduction

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Coplanar waveguide (CPW) has been finding wide applications in MIC and MMIC. It is more popular in millimeter-wave frequency due to several advantages it offers over microstrip line [1, 28], in particular, easiness in mounting series/shunt elements without needing via holes and low frequency dispersion of its dominant CPW mode. In the past few decades, great effort has been made to accurate characterization of a variety of CPW circuits and discontinuities by developing different advanced fullwave approaches [28, 87, 94]. Also, several commercial MoM simulators [48-50] have been explored in the recent years for direct simulation and optimization of a large number of CPW circuit blocks with complicated configurations. But, as detailed in [52], such a direct approach requires the intensive CPU time and huge CPU memory while it obstacles one from physical insight into the operation principle of the whole CPW circuit block. Therefore, it is preferred to formulate the equivalent circuit models for many CPW elements such that the optimization can be executed via resultant network topology.

In parallel, as one of key circuit blocks, CPW bandpass filter has been extensively studied towards realization of bandwidth enhancement, harmonic suppression and size-compactness as discussed in [84, 87, 97-100]. In [17, 84], the lumped capacitive and inductive gap coupling elements were originated to build up the initial end-coupled CPW filters using the half-wavelength ($\lambda / 2$) CPW resonators. Recent work has been shifted to explore the compact $\lambda / 4$ CPW filters with good harmonic suppression [97, 101]. But, due to their very limited coupling degree of these two elements, an alternative parallel-coupled CPW element was initiated in [97] and then employed in the filter design [101]. In [101], its coupling behavior was intuitively characterized in terms of 2-D even- and

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odd-mode characteristic impedances without including the dynamic effects caused by 3-D CPW discontinuity.

The purpose of this chapter is to derive the J- or K-inverter network parameters of the above lumped and distributed CPW coupling elements for comparative investigation of their coupling characteristics and further apply them for efficient filter design via circuit network topology. The developed “short-open calibration” (SOC) technique [52] has been extended to characterize the three CPW coupling elements in terms of their equivalent circuit networks. In this way, the asymmetrical parallel-coupled CPW with finite coupling length is modeled as a frequency-dispersive K-inverter network and it is then utilized as an inductive coupling element at both input and output ports in order to build up a size-compact and broadband $\lambda/4$ CPW bandpass filter. The two- and four-stage CPW bandpass filters are manually optimally designed and the predicted results are confirmed by those from the ADS software and fabrication measurement over the wide frequency range of 2.0 to 18.0 GHz.

4.2 MoM-SOC Modeling Technique

Figure 4.1 (a) describes the physical layout for modeling a two-port CPW gap coupling element using the fullwave MoM algorithm. As explained in chapter 3, the field over the slot area is perceived as a pair of equivalent magnetic current density below and above the perfect electric conductor according to the image principle. Plus, a pair of impressed electric current density, J_1 and J_2 , is simultaneously introduced at the two external ports

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selected far away from the gap discontinuity area so as to establish a set of source-type magnetic field integral equations (MFIE). Here, each feeding line length is chosen larger than 3.5mm which is enough for considered frequency range (2GHz-20GHz). Then, the magnetic current density over the slot area is expanded in terms of superposition of basis functions with the configuration of piecewise-sinusoidal and pulse functions as drawn in Figure 4.1(a). Moreover, according to the Galerkin's technique, a set of weighting functions with the identical number of basis functions are selected so as to convert the MFIE into a determinant matrix equation for numerical implementation, thus deriving the network parameters at the two external ports.

Figure 4.1 Physical layout and equivalent circuit network for fullwave MoM modeling of two-port CPW discontinuity including the configuration of basis-function and impressed current density at the ports. (a) layout; (b) circuit network.

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Of course, such an impressed field excites all the even higher-order modes in addition to our interested CPW dominant mode, but the former modes appear to rapidly decay and eventually disappear when they arrive at the two CPW open-ends [93]. Following our previous works [52], the whole layout in Figure 4.1(a) can be readily classified into two distinct parts, i.e., error terms and core network, in conjunction with the two feeding line sections and the circuit discontinuity under modeling as denoted in Figure 4.1(b), respectively. In this way, the error terms take into account all the effects caused by the impressed field at the ports and they are separately linked together with the central circuit network at the two sides.

Now, we move to employ the short-open calibration (SOC) technique to independently evaluate the electrical behavior of the above two error terms via characterization of two perfect calibration standards, i.e., short and open circuits in the consistent MoM platform. Figure 2.3(a) and (b) represent the physical formulation of these two standards under the odd- or even-symmetrical excitation of a uniform CPW with the length twice its corresponding feeding line, i.e., $2L_i$. So, the central symmetrical plane indicates the perfectly electric or magnetic wall and it can be considered as the ideal short- or open-end in the equivalent circuit. As denoted in Figure 2.3(a) and (b), half an overall structure exactly expresses the error term of our concern, which is terminated by the short- or open-end. Similarly to [52], such an error term can be numerically characterized via MoM algorithm and it can be simply expressed in terms of the three MoM-derived equivalent voltages at the source port for odd- and even-excitations as well as the open-end, i.e., V_{io} , V_{is} and V_{io}' , respectively. Moreover, the ABCD matrix $[X_i]$ of this error

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term can be explicitly derived as below,

$$[X_i] = \begin{bmatrix} \bar{V}_{io}' / (\bar{V}_{io} - \bar{V}_{is}) & -\bar{V}_{io}' / (\bar{V}_{io} - \bar{V}_{is}) \\ -1 / \bar{V}_{io}' & \bar{V}_{io} / \bar{V}_{io}' \end{bmatrix} \quad (4.1)$$

in which $\bar{V}_{io}$, $\bar{V}_{is}$, $\bar{V}_{io}'$ represent the three above-described voltages with respect to the equivalent impressed port current. As a consequence, the core circuit network parameters at the center in Figure 4.1(b) can be obtained by calibrating out the two error terms from those network parameters at the two ports based on a simple cascaded network theorem.

4.3 J- and K-Inverter Networks of CPW Coupling Structures

In this section, our particular attention is focused on fullwave characterization of a variety of CPW coupling structures. Based on the above description, the initial circuit models of these CPW elements can be numerically de-embedded via MoM-SOC technique and they can be directly expressed in terms of either Z- or Y-matrix. But, in order to characterize the capacitive or inductive coupling degree of these structures in a quantitative manner, it is always preferable to derive the relevant admittance (J-) or impedance (K-) inverter network.

4.3.1 J-inverter Network Topology

Figure 4.2 Two different sets of equivalent circuit networks for characterizing the CPW capacitive coupling structure. (a) π -type Y-matrix; (b) J-inverter network.

An admittance (J-) inverter is an ideal quarter-wave transformer or 90° phase shifter. If a series element Z_s with an admittance inverter with J value at either side, the input impedance can be expressed as:

$$Y_{in} = \frac{J^2}{Y_s}$$

Thus, the series element with the two admittance inverters will appear from the terminals to be equivalent shunt admittance. A π type circuits as shown in Figure 4.2 (a) actually can be converted to the admittance inverter with magnitude of J and two electrical lengths

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$\phi_1/2$ and $\phi_2/2$ at both sides as illustrated in Figure 4.2 (b). The negative element in J-inverter can be absorbed in to the elements of the adjacent resonant circuit to eliminate them from the overall network framework. Thus, the resultant filter can be then considered as circuits coupled by single susceptive elements.

The relationship between the parameters of these networks can be obtained by equivalence of ABCD matrix of two networks and has been explicitly established in [104]. The ABCD matrix of π type network is:

$$\begin{bmatrix} \frac{-B_{22}}{B_{12}} & j\frac{1}{B_{12}} \\ j\frac{B_{12}^2 - B_{11}B_{22}}{B_{12}} & -\frac{B_{11}}{B_{12}} \end{bmatrix} \quad (4.2)$$

And the ABCD matrix parameters of equivalent J-type network can be expressed as:

$$\begin{bmatrix} \cos \phi_1/2 & jZ_1 \sin \phi_1/2 \\ jY_1 \sin \phi_1/2 & \cos \phi_1/2 \end{bmatrix} \cdot \begin{bmatrix} 0 & 1/jB \\ -jB & 0 \end{bmatrix} \cdot \begin{bmatrix} \cos \phi_2/2 & jZ_2 \sin \phi_2/2 \\ jY_2 \sin \phi_2/2 & \cos \phi_2/2 \end{bmatrix}$$

It can be further simplified as:

$$\begin{bmatrix} \frac{B}{Y_1} \sin \frac{\phi_1}{2} \sin \frac{\phi_2}{2} + \frac{Y_2}{B} \cos \frac{\phi_1}{2} \sin \frac{\phi_2}{2} & j(\frac{B}{Y_1 Y_2} \sin \frac{\phi_1}{2} \sin \frac{\phi_2}{2} - \frac{1}{B} \cos \frac{\phi_1}{2} \cos \frac{\phi_2}{2}) \\ j(\frac{Y_1 Y_2}{B} \sin \frac{\phi_1}{2} \sin \frac{\phi_2}{2} - B \cos \frac{\phi_1}{2} \sin \frac{\phi_2}{2}) & \frac{B}{Y_2} \cos \frac{\phi_1}{2} \sin \frac{\phi_2}{2} + \frac{Y_1}{B} \sin \frac{\phi_1}{2} \cos \frac{\phi_2}{2} \end{bmatrix} \quad (4.3)$$

By equivalence of equation (4.2) and (4.3), the J-inverter network parameters are deduced as:

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$$\frac{J}{\sqrt{Y_1 Y_2}} = -\frac{\bar{B}_{12} \cos \frac{\phi_1}{2}}{\cos \frac{\phi_2}{2} + \bar{B}_{22} \sin \frac{\phi_2}{2}} \quad (4.4a)$$

$$\phi_1 = \arctan \frac{2(\bar{B}_{11} + \bar{B}_{22} |\bar{B}|)}{(1 + \bar{B}_{22}^2 - \bar{B}_{11}^2 - |\bar{B}|^2)} + m\pi$$

$$\phi_2 = \arctan \frac{2(\bar{B}_{22} + \bar{B}_{11} |\bar{B}|)}{(1 + \bar{B}_{11}^2 - \bar{B}_{22}^2 - |\bar{B}|^2)} + n\pi \quad (4.4b)$$

where $\bar{B}_{11} = B_{11} / Y_1$, $\bar{B}_{22} = B_{22} / Y_2$, $\bar{B}_{12} = B_{12} / \sqrt{Y_1 Y_2}$, and $|\bar{B}| = \bar{B}_{11} \bar{B}_{22} - \bar{B}_{12}^2$. Y_1 and Y_2 indicate the characteristic admittances of the two CPW feed lines, while the m and n are the integers to be determined.

4.3.2 K-inverter Network Topology

Similar to admittance inverter, an impedance inverter is also an ideal quarter-wave transformer or 90° phase shifter. Load impedance connected at one end can be seen as impedance that has been inverted with respect to the characteristic impedance squared at the input. If considering the parallel admittance element Y_s with an ideal impedance inverter with value K connected on both sides, the input impedance can be given:

$$Z_{in} = \frac{K^2}{Z_s}$$

Thus the shunt element with two impedance inverters converts the shunt admittance into equivalent series impedance.

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Figure 4.3 Two different sets of equivalent circuit networks for characterizing the CPW inductive coupling structure. (a) T-type Z-matrix; (b) K-inverter network.

Usually, CPW inductive coupling structure equivalent to a T-type impedance network shown in Figure 4.3 (a) can be converted to K-inverter network with K value and two different electrical lengths, $\phi_1/2$ and $\phi_2/2$ at both sides as illustrated in Figure 4.3 (b). According to the equivalence ABCD matrix of two networks, the corresponding K-inverter parameters can be derived.

The ABCD matrix of T-type network in Figure 4.3 (a) is:

$$\begin{bmatrix} \frac{X_{22}}{X_{12}} & j \frac{X_{11}X_{22} - X_{12}^2}{X_{12}} \\ -j \frac{1}{X_{12}} & -\frac{X_{22}}{X_{12}} \end{bmatrix} \quad (4.5)$$

while the ABCD matrix of K-inverter topology can be expressed as:

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$$\begin{bmatrix} \cos \phi_1 / 2 & jZ_1 \sin \phi_1 / 2 \\ jY_1 \sin \phi_1 / 2 & \cos \phi_1 / 2 \end{bmatrix} \cdot \begin{bmatrix} 0 & -jX \\ -j/X & 0 \end{bmatrix} \cdot \begin{bmatrix} \cos \phi_2 / 2 & jZ_2 \sin \phi_2 / 2 \\ jY_2 \sin \phi_2 / 2 & \cos \phi_2 / 2 \end{bmatrix}$$

It can be further simplified as:

$$\begin{bmatrix} \frac{Z_1}{X} \sin \frac{\phi_1}{2} \sin \frac{\phi_2}{2} + XY_2 \cos \frac{\phi_1}{2} \sin \frac{\phi_2}{2} & j(\frac{Z_1 Z_2}{X} \sin \frac{\phi_1}{2} \sin \frac{\phi_2}{2} - X \cos \frac{\phi_1}{2} \cos \frac{\phi_2}{2}) \\ j(XY_1 Y_2 \sin \frac{\phi_1}{2} \sin \frac{\phi_2}{2} - \frac{1}{X} \cos \frac{\phi_1}{2} \sin \frac{\phi_2}{2}) & \frac{Z_2}{X} \cos \frac{\phi_1}{2} \sin \frac{\phi_2}{2} + XY_1 \sin \frac{\phi_1}{2} \cos \frac{\phi_2}{2} \end{bmatrix} \quad (4.6)$$

By equivalence of equation (4.5) and (4.6), the K-inverter network parameters are deduced as [106]:

$$\frac{K}{\sqrt{Z_1 Z_2}} = \frac{-\sin \frac{\phi_1}{2} + \bar{X}_{11} \cos \frac{\phi_1}{2}}{\bar{X}_{12} \sin \frac{\phi_2}{2}} \quad (4.7a)$$

$$\phi_1 = \arctan \frac{2(\bar{X}_{11} + \bar{X}_{22} |\bar{X}|)}{(1 + \bar{X}_{22}^2 - \bar{X}_{11}^2 - |\bar{X}|^2)} + m\pi$$

$$\phi_2 = \arctan \frac{2(\bar{X}_{22} + \bar{X}_{11} |\bar{X}|)}{(1 + \bar{X}_{11}^2 - \bar{X}_{22}^2 - |\bar{X}|^2)} + n\pi \quad (4.7b)$$

where $\bar{X}_{11} = X_{11}/Z_1$, $\bar{X}_{22} = X_{22}/Z_2$, $\bar{X}_{12} = X_{12}/\sqrt{Z_1 Z_2}$, and $|\bar{X}| = \bar{X}_{11} \bar{X}_{22} - \bar{X}_{12}^2$. Z_1 &

Z_2 indicate the characteristic impedances of the two CPW while m and n are the integer numbers.

4.3.3 Extracted J- or K-inverter parameters

Figure 4.4 Geometry of the three CPW coupling structures to be characterized. (a) inductive gap CPW; (b) parallel-overlapped CPW; (c) capacitive gap CPW.

Figure 4.4(a) and (c) depict the layout of the two lumped inductive and capacitive coupling elements to be used in this chapter, respectively. They were originated in [17, 84] and successfully employed to develop the planar $\lambda/2$ [17, 84] and $\lambda/4$ CPW bandpass filters [99, 101]. Due to their weak coupling degree especially at low frequency, these elements can only be utilized to realize the narrow pass bandwidth. To effectively increment the concerned coupling degree, an alternative CPW parallel-coupled or overlapped structure was proposed in [97] and then employed to design a compact CPW

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filter [98]. Figure 4.4(b) shows its layout in which the two CPW lines with short-end are aligned in parallel with the overlapped length of T at the interval of a metallic gap.

Actually, in Chapter 3, these inductive or capacitive coupling elements have been modeled as their equivalent T- or π type network via MoM-SOC technique. Following the procedures described in previous sections, their corresponding equivalent K- or J-inverter parameters can be numerically obtained.

In the following, the above-described three coupling elements are realized on the substrate of RT6010: $\epsilon_r=10.8$ (permittivity) & $h=25\text{mil}$ (thickness), and the relevant results are plotted in Figure 4.5. Figure 4.5(a) describes the extracted normalized K-inverter impedance ($\bar{K}$) versus frequency for the CPW inductive coupling element, as shown in Figure 4.4(a), under the three varied gap widths (G). The normalized $\bar{K}$ is observed to rise up as a linear function of frequency regardless of different G. Further, its positive slope gradually increases as t is reduced from 15 to 5mil, thereby exhibiting the inductive coupling enhancement. From Figure 4.5(b), it can be also seen that the normalized J-inverter admittance ($\bar{J}$) of the lumped CPW capacitive coupling element in Figure 4.4(c) linearly increases with the frequency and gradually moves up as an incremental slope as G becomes narrow.

Figure 4.5(c) depicts the normalized $\bar{K}$ of the parallel-overlapped structure with inductive coupling over the wide frequency range of 4.0 to 18.0GHz. At low frequency, $\bar{K}$ is almost kept as a linearly incremental function of frequency while its linear scope is

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gradually raised as the overlapped length (T) is stretched from 30 to 100mil. But, as the frequency further increases, $\bar{K}$ at first rises up in a decelerating manner to achieve the

(a)

(b)

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(c)

Figure 4.5 Normalized K-inverter impedance ($\bar{K} = K / \sqrt{Z_1 Z_2}$) or J-inverter admittance ($\bar{J} = J / \sqrt{Y_1 Y_2}$) versus frequency under different coupling dimensions.

(a) inductive gap CPW; (b) capacitive gap CPW; (c) parallel-overlapped CPW.

relevant maximum value, 0.37 at 10.0GHz for the longest $T=100\text{mm}$, and then gradually falls down, as observed in Figure 4.5(c). Following our previous work in [106], the maximum coupling $\bar{K}$ appears at the frequency in which the overlapped T is approximately equal to the quarter-wavelength and further such a frequency keeps to be reduced as T is enlarged.

4.4 Model-Based CAD Design of CPW Filters

4.4.1 Wide-band bandpass filter

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Figure 4.6 Layout and equivalent cascaded network of a two-section $\lambda/4$ CPW bandpass filter as well as derived S-parameters. (a) layout; (b) equivalent network; (c) S-parameters

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With the use of the above-extracted circuit model, two quarter-wavelength ($\lambda/4$) CPW bandpass filters are efficiently designed via model-based network topology. Figure 4.6(a) shows the layout of a two-section CPW filter in which the parallel-overlapped coupling structures are put at both input and output ports due to its enhanced inductive coupling. Meanwhile, the two uniform CPW sections are linked together at the center through the CPW gap with capacitive coupling. Figure 4.6(b) indicates the relevant equivalent circuit network in which the electrical lengths at the two sides of K- or J-inverter, as in Figure 4.3(b) and Figure 4.3(a), are put together with the CPW section to make up the two $\lambda/4$ CPW resonators at the central frequency. Figure 4.6(c) is the model-based simulated S-parameters via the network in Figure 4.6(b) in comparison to those from the fullwave ADS-based simulation over the filter layout, in which $L_1=85\text{mil}$, $L_2=104\text{mil}$, $L_3=42\text{mil}$. They are very reasonably matched with each other and exhibit the two-pole passband behavior with the bandwidth of about 20% at the central frequency of 5.8GHz. But, slight discrepancy between them can be observed especially in the passband, and it is mainly caused by the under- or over-estimated K- inverter impedance of the overlapped-coupling due to different numerical discretization of the field distribution over the tight-coupled slots.

(a)

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(b)

Figure 4.7 Layout and equivalent cascaded network of a four-section $\lambda/4$ CPW bandpass filter. (a) layout; (b) equivalent network

Figure 4.7(a) shows the layout of the four-section $\lambda/4$ CPW bandpass filter. Herein, the two additional CPW sections are introduced and coupled with each other at the center through the inductive gap with weakly coupling degree. Similarly, such a four-section filter with large layout dimension can be optimally designed by carefully adjusting the corresponding parameters via its equivalent circuit network in Figure 4.7(b). Figure 4.8(a) and (b) show the model-simulated S-parameters against those from the ADS in the two different frequency scales. In this optimized case, the sizes of the three coupling structures are found: $L_4=100\text{mil}$, $L_5=47\text{mil}$, $L_6=5\text{mil}$, $L_7=88\text{mil}$, $L_8=202\text{mil}$, respectively. In Figure 4.8(a), the model-based results demonstrate the four transmission poles with the bandwidth of about 18% as compared with those from the ADS-based fullwave simulation of the overall layout at cost of much longer CPU time and large CPU memory. Again, the visible difference between them within the passband still exists due to the above-interpreted reasons. In this way, any small numerical approximation in the derived K- or J-inverter network parameters of each coupling element definitely causes the significant variation in the bandpass behaviors within the passband, such as disappearance of some transmission poles and degradation of the return loss, as also

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(a)

(b)

Figure 4.8 Derived S-parameters of a four-section CPW filter in different frequency ranges. (a) zoom-in scale; (b) zoom-out scale.

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found in the practical design of various planar multi-pole bandpass filters, e.g., [102]. But, both two sets of results are still found in a reasonable agreement with each other. Both sets of the results in Figure 4.8(b) in a large scale illustrate us that the 1st harmonic appears around the frequency of 16GHz, thereby comprising an ultra-stopband at the bandgap frequency with the range of 6.4GHz to 15.9GHz, i.e., bandwidth of 90%.

4.4.2 Harmonic suppressed $\lambda/4$ bandpass filter

In this section, asymmetric parallel-coupled coplanar waveguide (CPW) is characterized as an equivalent unified K-inverter network according to the procedures in Sec.4.3.2. The relevant network parameters are extracted to explicitly demonstrate the frequency-dependent coupling behavior with the 1st zero coupling at the $\lambda/2$ length of the overlapped coupling stage. By differentiating the coupled spaces and overlapped lengths, the two coupling structures are dissimilarly constructed and utilized as the input/output excitation of a CPW bandpass filter. Through allocating the two unequal zeros towards the spurious harmonic around $3\lambda/4$, an improved two-stage $\lambda/4$ CPW filter is designed at 5.0GHz. Optimized results achieve broad rejection band, ranging from 5.8GHz beyond 18.0GHz, as verified by the ADS simulation. Following the similar ideal, a four-stage $\lambda/4$ CPW bandpass filter ended with two asymmetric parallel-coupled stages is proposed and its performances are verified by experimental results.

In Sec.4.4.1, the parallel-coupled CPW element and traditional lumped capacitive or inductive coupling element has been successfully utilized to design wide band pass filter

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with size-compactness and 1st harmonic suppression by stagger excitation of $\lambda/4$ CPW resonators with inductive and capacitive coupling elements.

In this section, we also characterize this distributed CPW coupling structures into its equivalent K-inverter network via our self-calibrated method of moments, namely, MoM-SOC. Extensive results demonstrate that the relevant coupling degree, i.e., K-impedance, varies as a quasi-sinusoidal function of frequency while its value becomes zero at the frequency with the $\lambda/2$ overlapped length. With the use of the two dissimilar CPW coupling elements at the input/output ports, an improved $\lambda/4$ CPW filter with broad upper rejection band is designed by allocating the two unequal coupling zeros to the 2nd harmonic.

The two CPW short-ends with different strip widths (w_1 and w_2) are overlapped in a parallel way as indicated in Figure 4.4 (b) to make up a distributed CPW inductive coupling structure. Following the work in Sec.4.3.3, this asymmetric CPW structure with two dissimilar CPW feeders can be accordingly expressed as an equivalent K-inverter network with the K-impedance (K) and two unequal electrical lengths at its two sides ($\phi_1/2$ and $\phi_2/2$).

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(a)

(b)

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Figure 4.9 K-inverter network parameters of the parallel-overlapped CPW structure.

(a).(c).(d).normalized K-impedance; (b) equivalent electrical length

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Figure 4.9 (a) depicts the extracted K-inverter impedance versus overlapped length (T) as a function of frequency (f). As frequency increases, the K-impedance is observed to gradually go up to its maximum and then fall down across the zero point, thereby exhibiting the quasi-sinusoidal frequency dispersion. Furthermore, we can find that this transmission zero is moved down as the overlapped length (T) is extended. Figure 4.9 (b) shows the two relevant electrical lengths versus frequency. It can be seen here that the K-impedance basically achieves its maximum around the frequency where $\phi_2/2=90^\circ$ while gets to the zero minimum when $\phi_2/2=180^\circ$. Figure 4.9(c) describes the normalized K-impedance versus spacing (d) between the overlapped-CPW slots. K-impedance curve tends to decrease with d since the coupling between the two CPW is reduced. In order to construct a $\lambda/4$ CPW filter with harmonic suppression in the next section, the two dissimilar CPW coupling structures with narrow and wide spaces, i.e., d_1 and d_2 , are further characterized to achieve the unchanged coupling degree at the central frequency (f_0) of the dominant passband while allocating the two transmission zeros to f_1 and f_2 , respectively, around the 2nd harmonic passband. It can be realized by extending the CPW overlapped length with d_2 as illustrated in Figure 4.9 (d).

Figure 4.10(a) depicts the proposed $\lambda/4$ CPW bandpass filter. The two $\lambda/4$ CPW resonators with unequal strip/slot widths are inductively excited by the two parallel-coupled CPW with different spaces (d_1 and d_2) and overlapped length (L_1 and L_5) while they are capacitively coupled together via gap width (L_3). Figure 4.10(b) indicates the relevant equivalent cascaded topology with the pre-calculated two K-impedance (K_1 , K_2) and one J-susceptance (J) via MoM-SOC as well as the two $\lambda/4$ transmission line

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resonators at the central frequency.

Figure 4.10 The proposed two-stage CPW $\lambda/4$ bandpass filter. (a) physical layout; (b) equivalent circuit topology.

Figure 4.11(a) illustrates the three sets of calculated insertion losses (S_{21}) in conjunction with the three cases. For the traditional $\lambda/4$ CPW filter without adjusting the overlapped length [101] at both CPW coupling elements, the spurious harmonic appears around 13.6 to 16.8GHz with two peaks as observed in the solid curve. By suitably adjusting the overlapped length to $L_1=L_5=147\text{mil}$ under the other dimensions of $d_1=d_2=7\text{mil}$, $L_2=L_4=80\text{mil}$ and $L_3=28\text{mil}$, the single transmission zero can be observed and it is suitably allocated to 13.8GHz, thereby canceling the 1st peak as shown in the dashed curve. Following the description in Figure 4.9(d), the two asymmetric CPW structures

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(a)

(b)

Figure 4.11 Derived S-parameters of the proposed $\lambda/4$ CPW bandpass filters. (a) three sets of results for physical understanding; (b) optimized results with ADS simulation

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with $d_1=5\text{mil}$, $d_2=7\text{mil}$, $L_1=126\text{mil}$ and $L_5=147\text{mil}$ are formulated to simultaneously bring out the two transmission zeros targeting the two separate peaks, i.e., 13.8GHz and 16.2GHz. As exhibited in the dotted line in Figure 4.11(a), both peaks are consequently canceled out so as to effectively suppress this spurious harmonic passband.

Figure 4.11(b) depicts the optimized results as compared to the direct ADS simulation. Filter specifications are as following: $f_0=5.0\text{GHz}$ and $\text{BW}=28\%$; substrate: $\epsilon_r=10.8$ and $h=25\text{mil}$; dimensions: $d_1=5\text{mil}$, $d_2=7\text{mil}$, $L_1=126\text{mil}$, $L_2=99\text{mil}$, $L_3=25\text{mil}$, $L_4=80\text{mil}$ and $L_5=147\text{mil}$. Both curves illustrate the two transmission poles within the dominant passband of the fractional bandwidth of $\text{BW}=28\%$ at the central frequency of $f_0=5.0\text{GHz}$. Meanwhile, the upper rejection band is significantly broadened to the range from 5.8GHz beyond 18.0GHz relying on the separate allocation of the two transmission zeros at the frequencies of 14.0GHz and 16.6GHz. Thus, as discussion above, an effective method for spurious harmonic suppression in the CPW bandpass filter design by using the two asymmetric parallel-coupled CPW at the input/output ports. Such a distributed coupling structure is modeled as a unified K-inverter network so as to quantitatively demonstrate its frequency-dependent coupling behavior in a quasi-sinusoidal-function variation. Detailed discussion is made to exhibit its transmission zero as a function of overlapped length and further allocate the two zeros to suppress the 2nd spurious harmonic. Optimized results on a proposed two-stage $\lambda/4$ CPW bandpass filter show us an ultra-broad rejection band with simultaneous cancellation of the 1st, 2nd and harmonic passbands as confirmed by the ADS-simulated results. The large return loss at $f=14\text{GHz}$ in ADS simulation is caused by radiation, however, this issue is not considered in

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network based method.

To further improve the stopband performance, a four-pole $\lambda/4$ coplanar waveguide (CPW) bandpass filter, ended with the two asymmetric parallel-coupled stages following the above ideal [106], is proposed. By separately allocating the two unequal transmission zeros, the spurious response at the triplicate passband frequency ($3f_0$) is suppressed. Simulated and measured results confirm the attractive features of this proposed filter, i.e., wide passband and good harmonic suppression.

The topology of the proposed four-pole $\lambda/4$ CPW bandpass filter is shown in Figure 4.12 and it is designed on the RT/Duriod 6010 substrate with $\epsilon_r=10.8$ and $h=25\text{mil}$. In the middle section, the lumped capacitive and inductive coupling elements are used for cascaded connection of two neighbored $\lambda/4$ CPW resonators. At the two ends, the two asymmetric parallel-coupled CPW stages are constructed to realize the tightened coupling degree as always needed in the filter design. These two CPW stages are made up with unequal coupling gap widths, d_1 & d_2 , and overlapped lengths, L_1 and L_2 , as shown in Figure 4.12. Figure 4.13 illustrates the normalized K-inverter impedance ($\bar{K} = K/\sqrt{Z_{01}Z_{02}}$) of these two coupled CPW stages as a function of frequency. As can be seen, both normalized $\bar{K}$ has the almost same value at the passband frequency ($f_0=4.6\text{GHz}$) such that the identical coupling degree is achieved at two coupling ends. Further, the two $\bar{K}$ values become unequal as the frequency increases beyond f_0 , thus generating the two separated transmission zeros, f_1 & f_2 , around $3f_0$. These two zeros play a critically

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important role in effectively canceling the 1st-harmonic passband at $3f_0$ of this proposed $\lambda/4$ CPW filter.

Figure 4.12 Topology of the proposed four-pole $\lambda/4$ CPW bandpass filter

Figure 4.13 Normalized K-inverter impedance of the parallel-coupled CPW with different gap widths (d) and overlapped lengths (L)

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(a)

(b)

Figure 4.14 Predicted and measured S-parameters of the proposed CPW filter. (a) wide frequency range (b) narrow frequency range

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Figure 4.14(a) illustrates the predicted and measured S-parameters of the optimized CPW filter over the wide frequency range of 2 to 18GHz. In this design, the two asymmetric coupled CPW sections with the dimensions: $d_1=5\text{mil}$, $d_2=7\text{mil}$, $L_1=130\text{mil}$ and $L_2=147\text{mil}$, are formed to produce the two transmission zeros at 13.7 and 15.6GHz, respectively. As such, the S_{21} -magnitude in the whole spurious passband with the region of 13.0 to 16.0GHz is reduced below or close to -10dB as can be observed from the simulated and measured curves in Figure 4.14 (a). Figure 4.14 (b) shows the narrow-range frequency dispersion of these S-parameters. Simulated and measured results seem to agree well with each other except about 4.0% frequency shift. The fractional bandwidth (BW) is calculated as about 26.0% as compared with a few percentages of the CPW filters using lumped coupling elements [101]. Within the passband of 4.0 to 5.2GHz, the return loss and insertion loss in experiment achieve higher than 13dB and lower than 1.5dB, respectively.

In conclusion, a novel four-pole $\lambda/4$ CPW bandpass filter with wide passband and harmonic suppression is presented using the two asymmetric parallel-coupled CPW stages at the input/output ends. After the principle of exciting the two transmission zeros is discussed based on the K-inverter of a parallel-coupled CPW, such a CPW filter is optimally designed and its predicted performance is confirmed by experiment.

4.5. Conclusion

In this chapter, a class of compact quarter-wavelength bandpass filters is proposed,

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designed and implemented. First, the two traditional lumped CPW inductive/capacitive coupling elements and the distributed parallel-overlapped coupling structure are characterized via self-calibrated MoM-SOC technique. All of them are further modeled as either K- or J-inverter network so as to explicitly demonstrate their frequency-dispersive coupling behaviors in comparison. Next, model-based optimization design of multi-section CPW bandpass filters is carried out via cascaded network topology. By manually adjusting each dimension of coupling elements and transmission lines' lengths, two- and four-section $\lambda/4$ CPW bandpass filters can be easily designed via extracted model parameters. The derived S-parameters are reasonably verified by the ADS-based approach. The one-round computational time is less than 20 seconds as compared to 6 hours when the simulation is executed via ADS-Momentum under the selection of 20 cells per wavelength at 20GHz. The simulated results exhibit several attractive features of these filters, i.e., broader pass bandwidth of about 20% and wider/sharper upper stopband with the bandwidth of 90%.

Also, the performance of stopband is further improved by introducing two transmission zeros through parallel-coupled element. Such a distributed coupling structure is modeled as a unified K-inverter network so as to quantitatively demonstrate its frequency-dependent coupling behavior in a quasi-sinusoidal-function variation. Detailed discussion is made to exhibit its transmission zero as a function of overlapped length. By carefully allocating these two zeros, the 2nd spurious harmonic can be suppressed. Optimized results on a proposed two-stage $\lambda/4$ CPW bandpass filter show us an ultra-broad rejection

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band with simultaneous cancellation of the 1st, 2nd harmonic passbands as confirmed by the ADS-simulated results.

Finally, a novel four-pole $\lambda/4$ CPW bandpass filter with wide passband and harmonic suppression is presented. The performances in passband and stopband are confirmed by experiment results.

Chapter 5

Ultra Wideband (UWB) Bandpass Filters

Abstract: In this chapter, two novel ultra-wideband (UWB: 3.1~10.6GHz) bandpass filters on coplanar waveguide (CPW) are presented, designed and implemented. At first, open-ended/short-ended non-uniform or multiple-mode resonator with three distinctive sections is constructed towards generating the first three resonant modes occurring around the lower-end, center and higher-end of the UWB band. Then, a CPW interdigital capacitor/paralleled-coupled inductor with enlarged ground-to-ground distance is characterized to excite two additional resonant poles below and above the UWB's center. As a result, two five-pole UWB bandpass filters with only one full-wavelength are constituted. Their performances are studied on the basis of a simple cascaded transmission-line network, whose parameters are extracted from our self-calibrated method of moments. After the optimized results are confirmed by a full-wave simulation over the whole filter layout, two UWB filter samples are fabricated to demonstrate the

actual UWB passband behavior with the bandwidth (110%) as well as insertion loss ($<1\text{dB}$) and return loss ($>10\text{dB}$).

5.1 Introduction

In February 2002, the Federal Communications Commission (FCC) in USA released the unlicensed use of ultra-wideband (UWB) wireless systems in the 3.1 to 10.6GHz frequency band [107]. Since then, much research interests and activities have been aroused on exploring a variety of UWB systems [108] as well as UWB components and devices [109]. Of them, the UWB bandpass filters have been considered as one of the key circuit blocks in the whole UWB systems. However, as well known, the matured microwave filter theory [110] was in general established under the assumption of narrow passband. As such, this theory is not applicable for design of such a UWB filter, mainly due to the three factors: insufficiently high coupling degree in coupling elements, unexpected spurious resonances in quarter- or half-wavelength resonators, and uncontrollable non-linear frequency dispersion over the 110% UWB-band.

For this reason, extensive research works have recently been executed to propose and exploit a few types of wideband planar bandpass filters with varied configurations and approaches [111-115]. The fundamental target of these works is to constitute such a FCC-defined UWB passband with a fractional bandwidth of 110% at the central frequency of 6.85GHz. To our best knowledge, the two pioneered works in [111, 112] for the first time reported two initial UWB bandpass filters in 2003. In [111], a microstrip line is mounted in a lossy composite substrate so as to realize a high attenuation at high frequencies. In

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[112], a dual-stopband microstrip line ring filter is constructed to realize a UWB passband by properly allocating the two relevant attenuation poles below 3.1GHz and above 10.6GHz. Very recently, a composite UWB filter [113] is constituted by cascading lowpass and highpass filter structures together or embedding one into the other. Meanwhile, a surface-to-surface or broadside coupled microstrip-coplanar waveguide (CPW) structure with tightened coupling degree is utilized to design an alternative UWB filter with one, two and three sections [114]. Moreover, a novel compact UWB bandpass filter on microstrip line is reported in [115] on a basis of a single multiple-mode resonator (MMR) and frequency-dispersive parallel-coupled lines with enhanced degree. Its basic principle is originated in [105] and applied in [116] to achieve a broad fractional passband of 60%~ 80% in terms of the first two or three resonant modes in a non-uniform or stepped-impedance resonator [117].

On the other hand, research on coplanar waveguide (CPW) bandpass filters [17, 84, 97, 99, 118-119] has been attracting a substantial interest in the exploration of uniplanar microwave integrated circuits and systems. The two initial CPW filters are presented in [17, 84] by cascading multiple half-wavelength CPW resonators through capacitive or inductive end-coupling elements. An improved CPW filter with compact size and wide stopband is reported in [97] with the use of quarter-wavelength CPW resonator that is capacitively and inductively end-coupled at its two ends. Due to extremely weak coupling extent of these end-coupling elements, the resultant CPW filters usually suffer from narrow passband (less than or about a few percentages) and high insertion loss. To circumvent this issue, various CPW edge- or parallel-coupled structures are proposed [99]

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and applied [116] to make up a compact wideband CPW filter with the 50%~60% fractional bandwidth. Recently, a simple CPW bandpass filter based on the combination of CPW lowpass and highpass periodic structures is presented in [119] and its passband achieves about 60% bandwidth. However, no research work has been reported to successfully develop any CPW bandpass filter with the passband covering the entire UWB band (3.1 to 10.6GHz).

In this chapter, two novel UWB bandpass filters on coplanar waveguide (CPW), as illustrated in Figure 5.1 and Figure 5.9, are proposed, designed and implemented. To do it, open end or short end CPW multiple-mode resonator (MMR) [115] are characterized to allocate its first three resonant modes occurring near the lower-end, center and higher-end of the targeted UWB passband. Then, a CPW interdigital capacitor [118] with enlarged ground-to-ground edges or paralleled-coupled inductor [97] are investigated to excite the two additional poles [105, 116] below and above the UWB's center or 6.85GHz. Afterwards, 5-pole CPW wideband filters with the UWB passband are designed on a basis of simple equivalent transmission line network. In this aspect, all the network parameters are extracted over a wide frequency range from our fullwave method of moments (MoM) via short-open calibration (SOC), namely, MoM-SOC approach [52]. After our network-predicted performances are confirmed by our fullwave simulation of an overall filter layout, two CPW filters are fabricated to show the actual UWB passband behaviours with insertion loss $< 0.5\text{dB}$, return loss $> 10\text{dB}$ and group delay variation $< 0.23\text{ns}$ over the entire UWB passband.

5.2 Open-ended UWB Bandpass Filter

Figure 5.1 Geometry of the proposed UWB bandpass filter on coplanar waveguide.

5.2.1 Open-ended Multiple-Mode Resonator on CPW

Figure 5.2 (a) Geometry and (b) equivalent circuit network of the open-ended CPW multiple-mode resonator (MMR) to be investigated.

Figure 5.2(a) depicts the open-ended CPW stepped-impedance multiple-mode resonator (MMR) to be characterized. It is composed of three distinctive sections, one low-impedance section in middle and two identical high-impedance sections on two sides.

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Figure 5.2(b) is its equivalent transmission line network. As pointed out in [115], this proposed MMR aims to make effective use of its lowest multiple resonant modes in wideband filter design. Even though this MMR is very similar to the stepped-impedance resonator (SIR) [117] in geometry and equivalent topology, the SIR uses the only lowest or dominant resonant mode in the design of narrowband filters with widened upper stopband. To manifest its multiple-mode characteristics, this MMR is at first analyzed on a basis of a simple transmission line topology. As shown in Figure 5.2(b), parasitic effects of open-ends and step discontinuities are ignored in order to simplify our analysis. To establish the resonant condition for all the modes, let's first derive the input admittance (Y_{in}) at the left open-end, looking into the right side, as indicated in Figure 5.2 (b).

$$Y_{in} = jY_2 \cdot \frac{2(K \tan \theta_1 + \tan \theta_2) \cdot (K - \tan \theta_1 \tan \theta_2)}{K(1 - \tan^2 \theta_1) \cdot (1 - \tan^2 \theta_2) - 2(1 + K^2) \tan \theta_1 \tan \theta_2} \quad (5.1)$$

where $K = Y_1 / Y_2$ is the admittance ratio of this MMR. Under the resonance condition, $Y_{in} = 0$ is valid. As such, an algebraic equation is established to determine all the resonant frequencies, including the three lowest ones of interest, i.e., f_1 , f_2 and f_3 . In this design, the electrical lengths of three sections are readily selected as $\theta_1 \approx \theta_2 = \theta$ such that three closed-form equations can be deduced to separately determine f_1 , f_2 and f_3 .

$$\theta(f_1) = \arctan \sqrt{K} \quad (5.2a)$$

$$\theta(f_2) = \pi / 2 \quad (5.2b)$$

$$\theta(f_3) = \pi - \arctan \sqrt{K} \quad (5.2c)$$

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Figure 5.3 (a) Three lowest resonant frequencies versus CPW end-side slot width (s), and (b) Frequency responses of S_{21} -magnitudes under $s=10, 55, 100$ mil.

In this simplified case, the lower and higher frequencies, f_1 and f_3 , are mainly controlled by K , while the central f_2 relies on the selected actual lengths of three sections in this MMR.

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At the beginning, the central and two side sections are properly chosen close to the half and quarter guided-wavelength at the UWB's central frequency of 6.85GHz, $2L_1 = \lambda_{g1}/2$ and $L_2 = \lambda_{g2}/4$, respectively. In addition, the central CPW section is formed with the fixed $Z_1 = 50\Omega$, and it has the strip/slot widths of 50 and 15mil. Now, let's investigate the variation of these three frequencies as a function of the slot width (s) in the two side sections and the relevant results are plotted in Figure 5.3(a). As s increases, all the three resonant frequencies, f_1 , f_2 and f_3 , tend to monotonically rise up in terms of different incremental slopes. To illustrate the concerned multiple-mode resonances in a straightforward way, this MMR resonator is weakly coupled with two 50Ω CPW feed lines via capacitive gaps [17] and the relevant frequency responses of S_{21} -magnitude are plotted in Figure 5.3(b) under $s = 10, 55$ and 100mil . First of all, three peaks appear in the 3 to 11GHz frequency range and they are caused by the above-discussed three resonant modes. Then, the locations of these peaks are found to move downwards and/or upwards as s varies. According to the criteria in [115] that the three resonant modes should equally occur above the lower-end, close to the center and below the upper-end of 3.1 to 10.6GHz UWB passband, this slot width is selected close to $s = 55\text{mil}$ in our design of a UWB bandpass filter in the following. To do it, a dispersive-frequency coupling element with enhanced extent needs to be constructed to achieve a flat UWB passband with minimized insertion loss [115].

5.2.2 CPW Interdigital Capacitor

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Figure 5.4 (a) Geometry and (b) equivalent circuit model of an improved CPW interdigital capacitor with enhanced coupling degree.

Figure 5.4(a) describes the geometry of an improved CPW interdigital capacitor element with enlarged ground-to-ground distance. Its original CPW counterpart was investigated using the two-dimensional (2-D) static approach in [118] to show the relevant coupling performance under the unsymmetrical coupled-line theorem. In the analysis [118], all the parasitic effects, such as open-ended and step-connected discontinuities as well as frequency dispersion, are ignored. Also, its coupling properties are limitedly exhibited in a narrow frequency range in which the length of the finger or coupled-section can be assumed as one quarter-wavelength. But, over a 110% UWB passband that is the case considered in this work, such a CPW interdigital capacitor needs to be characterized as a frequency distributed CPW element. Follow the works in [103, 106], the overall CPW

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structure in Figure 5.4(a) is in general expressed as a cascaded network that consists of a J-inverter network in the middle and two feed lines in the two sides as shown in Figure 5.4(b).

In this context, the coupled section with finger configuration is represented by a frequency-dependent J-inverter susceptance (J) and two unequal electrical lengths (ϕ_1 and ϕ_2). Those network parameters can be numerically extracted by executing the short-open calibration (SOC) procedure in the fullwave method of moments (MoM) at the two reference planes, P_1 and P_2 , as detailed in Chapter 4. As such, the J-susceptance as well as two electrical lengths can be derived in terms of the three independent parameters of an unsymmetrical Y-matrix, B_{11} , $B_{12}=B_{21}$ and B_{22} [104].

$$\frac{J}{\sqrt{Y_1 Y_2}} = -\frac{\bar{B}_{12} \cos \phi_2}{\cos \phi_1 + \bar{B}_{11} \sin \phi_1} \quad (5.3a)$$

$$\tan(2\phi_1) = \frac{2(\bar{B}_{11} + \bar{B}_{22}|\bar{B}|)}{(1 + \bar{B}_{22}^2 - \bar{B}_{11}^2 - |\bar{B}|^2)} \quad (5.3b)$$

$$\tan(2\phi_2) = \frac{2(\bar{B}_{22} + \bar{B}_{11}|\bar{B}|)}{(1 + \bar{B}_{11}^2 - \bar{B}_{22}^2 - |\bar{B}|^2)} \quad (5.3c)$$

where $\bar{B}_{11} = B_{11} / Y_1$, $\bar{B}_{22} = B_{22} / Y_2$, $\bar{B}_{12} = B_{12} / \sqrt{Y_1 Y_2}$ and $|\bar{B}| = \bar{B}_{11} \bar{B}_{22} - \bar{B}_{12}^2$.

To provide a comprehensive explanation on the coupling performance of this CPW interdigital capacitor, the impedance mismatching between the CPW feed lines ($Z_0=1/Y_0$) and two dissimilar middle lines ($Z_1=Y_1/2$ and $Z_2=1/Y_2$) should be taken into account. As

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investigated in [105] for a microstrip UWB filter, this improved CPW coupling structure is expected to generate the two additional transmission poles for such a UWB filter design. These two poles can be intuitively explained in [116] by the excitation of two quarter-wavelength resonators which are capacitively coupled with each other and inductively coupled to their respective feed lines. Here, an alternative approach is presented in terms of the relevant J-inverter parameter itself to explain these two poles and their locations. On a basis of the filter theorem in [110], the two input admittances, Y_1^* and Y_2^* , at the two planes, #1 and #2, shown in Figure 5.4(b) can be actually referred as to the two image admittances for the J-susceptance. As such, the resultant coupling degree of this CPW capacitor structure can be quantitatively judged and represented in terms of the normalized J-susceptance with respect to the two-side image admittances, i.e., $\bar{J} = J / \sqrt{Y_1^* Y_2^*}$.

Now, we'd like to move to demonstrate this frequency- dependent normalized J-susceptance over a wide range under different dimensions involved in the coupled section. Figure 5.5(a) depicts the three curves of this susceptance in the cases of $s=20, 30, 40$ mil under the fixed length $d_3=150$ mil. Regardless of different cases, this $\bar{J}$ goes up at first to the peak and then fall down. As pointed in [105, 116], this peak emerges at the frequency where the two-side length are equal to the quarter wavelengths or $\phi_1 = \phi_2 = \pi / 2$. As the slot width (s) increases, the coupling peak gradually rises up from 0.82 to 1.08 and 1.24, indicating that the coupling between the parallel-coupled strip conductors in the central part receives a gradual enhancement. Based on the analysis in [105], we can figure out that the two frequencies with $\bar{J} = 1$ in the two sides of the peak actually create the two

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Figure 5.5 Frequency-dependent normalized J-susceptance under different (a) slot width (s), and (b) coupled length (d_3)

transmission poles with zero or minimum S_{21} -magnitude. In order to apply these two poles in the UWB filter design, Figure 5.5(b) is plotted to show us the variation of two poles, f_a and f_b , with respect to the coupled length, d_3 . They are both moved downwards

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as d_3 is stretched. With these exhibited five poles, i.e., f_a , f_b , f_1 , f_2 , f_3 in mind, a five-pole UWB bandpass filter on CPW can be realized.

5.2.3 UWB Bandpass Filter on CPW

Figure 5.6 (a) schematic layout and (b) equivalent cascaded network topology of the proposed UWB bandpass filter on CPW

Figure 5.1 is the three-dimensional (3-D) view of the proposed open-end UWB bandpass filter on CPW. This filter is composed of a single multiple-mode resonator (MMR) and two dispersive CPW interdigital capacitors. To achieve the specified UWB passband, the three sections of this MMR should be arranged to be about quarter-, half- and quarter-wavelength, i.e., $\lambda_{g2}/4$, $\lambda_{g1}/2$ and $\lambda_{g2}/4$, as indicated in Figure 5.1. Figure 5.6(a) depicts the top view of this UWB filter with all the known dimensions except a varied coupling length (d_3). Its equivalent network topology, shown in Figure 5.6(b), is utilized here to

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carry out a manual optimization design of such a UWB filter in terms of the network parameters reported in Sec. 5.2.1 and Sec. 5.2.2. Looking at Figure 5.6(b), we can figure out that the three cascaded transmission line sections in middle with the phases, ϕ_2 , ϕ_3 and ϕ_2 , comprise the resultant MMR.

In the design using the network in Figure 5.6(b), the dimensions of this MMR are at first determined as done in Sec.5.2.1. Then, the coupling length (d_3) is properly chosen around quarter-wavelength of 6.85GHz with the target of minimizing the insertion loss within the UWB passband. Figure 5.7(a) describes the frequency response of S_{21} -magnitude for the filter structure in Figure 5.6(a) as d_3 is stretched from 20 to 80 and 162mil (optimized case). In the case of $d_3=20$ mil, the three resonant frequencies with maximum S_{21} -magnitude emerge within the shaded UWB band and they are allocated at 4.2, 7.0 and 9.8GHz, respectively. As d_3 is raised to 80 and 162mil, the lower cut-off frequency is approaching 3.1GHz (lower-end of UWB band) and the S_{21} -magnitude in the UWB passband rises up close to 0-dB line. On the other hand, the S_{21} -magnitude in the upper range beyond the UWB band at first goes up to 0-dB line since the distributed coupling degree of CPW interdigital capacitor is maximized in this range. Afterwards, it quickly falls down as the coupling zero of this distributed CPW capacitor is lowered to this range [115]. To confirm the predicted electrical performance derived from the simple network in Figure 5.6(b), the overall optimized UWB filter with $d_3=162$ mil is simulated over its entire layout using the Agilent Momentum at the cost of long CPU time and large memory. This Momentum software is embedded into the Agilent ADS for fullwave simulation of various multi-layer transmission-line structures and circuits. For an instance,

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(a)

(b)

Figure 5.7 Simulated frequency response of S_{21} -magnitude of the UWB filter structure in Figure 5.6(a). (a) different coupled lengths (d_3), (b) comparison between the network-optimized and fullwave-simulated results.

the mesh setup in Momentum is selected as 20 cells per wavelength at 20GHz for one short-ended filter case, the one-round simulation is more than 6 hours and requires the

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memory of about 500M. The two sets of simulated results are plotted together in Figure 5.7(b) and they are found in good agreement with each other over the wide frequency range of 2.0 to 12.0GHz.

5.2.4 Measured UWB Passband Performance

Figure 5.8 Measured and simulated results of fabricated UWB bandpass filter (a) layout for fabrication (b) Magnitudes of S-parameters and group delay

In this part, the optimized UWB filter is fabricated and its layout is shown in Figure 5.8 (a). This five-pole UWB filter occupies the overall length of 735mil, i.e., about one guided-wavelength at 6.85GHz, indicating that its length is much shorter than those

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reported in [111-114]. Figure 5.8(b) shows the measured S-parameter magnitudes ($|S_{11}|$ and $|S_{21}|$) and group delay in the 1.0 to 13.0GHz frequency range. It can be found here that the initial measured results have almost realized the FCC-specified UWB passband, i.e., 3.1 to 10.6GHz, under 3-dB definition of cutoff frequencies. Within the UWB band, the maximum insertion loss and minimum return loss achieve 1.5dB and 7.0dB, respectively, as compared to 0.3dB and 12.0dB in the simulated results. Meanwhile, the lower and upper cutoff frequencies are found to be slightly shifted from 3.1 and 10.6GHz to 2.8 and 10.2GHz. All these minor discrepancies may be attributed to the three uncontrollable aspects. In the simulation, the strip conductor thickness of 0.7~1.4mil and its roughness in side edges can not be taken into account in the multilayer-targeted MoM software or simulator. As marked in the data sheet of the used RT/Duroid 6010, the actual permittivity of this substrate is not exactly equal to 10.8 and its value at 10GHz is given as 10.8 ± 0.25 . In addition, the unexpected impedance mismatching effects in the connector between coaxial cable and CPW feed line may affect the measured return/insertion losses in the UWB passband. Figure 5.8(b) also shows us that the measured group delay within the UWB passband varies from 0.25 to 0.58ns or its maximum variation achieves 0.33ns, thus indicating that the transmission-coefficient phase of this proposed UWB filter is a linear function of frequency as well.

5.3 Short-ended UWB Bandpass Filter

In this section, another UWB filter with short-ended MMR and paralleled-coupled or overlapped inductive coupling element is designed and implemented following the

similar design procedures of open-ended type.

Figure 5.9 Geometry of short-ended MMR UWB filter on CPW

Figure 5.9 shows the proposed 3-D geometry of short-ended UWB filter. Similar to the above discussed open-ended UWB filter, the short-ended MMR with three distinctive sections is firstly formed to generate three resonant modes around the lower-end, center and upper-end of the UWB band. Then overlapped inductive coupling element is characterized via K-inverter to excite two additional poles below and above the center frequency, thus another five-pole UWB filter is constituted and its performances are well verified by fabrication results.

5.3.1 Short-ended MMR

Figure 5.10 (a) depicts the short-ended CPW step-impedance multiple-mode resonator (MMR) which is composed of three distinctive sections: one high-impedance in the middle and two identical low-impedance at both sides. Figure 5.10 (b) is equivalent circuit network. As pointed out in Sec.5.2.1, three lowest multiple resonant modes will be utilized for wideband filter design.

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Figure 5.10 (a) Geometry and (b) equivalent circuit network of the short-ended CPW multiple-mode resonator (MMR)

In this case, the input impedance (Z_{in}) at the left side of short end, looking into the right side, can be derived as:

$$Z_{in} = jZ_2 \cdot \frac{2(K \tan \theta_1 + \tan \theta_2) \cdot (K - \tan \theta_1 \tan \theta_2)}{K(1 - \tan^2 \theta_1) \cdot (1 - \tan^2 \theta_2) - 2(1 + K^2) \tan \theta_1 \tan \theta_2} \quad (5.4)$$

Here, $K = Z_1 / Z_2$ is the impedance ratio of this MMR. The resonator condition is enforced by: $Z_{in} = 0$. In this filter design, electrical lengths are assumed to be $\theta_1 \approx \theta_2 = \theta$ thus these three resonant frequencies f_1, f_2, f_3 also satisfy the equations (5.2a), (5.2b), (5.2c).

Based on the discussion on the open-ended MMR, the first design step is to approximately locate three resonator frequencies at around 4.2 GHz, 7.0GHz, and 9.8 GHz. Thus the impedance ratio K is chosen around 1.85. In order to simplify the analysis,

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the high-impedance section of MMR is around $50\ \Omega$ with half-wavelength $\lambda_{g1}/2$ at UWB's central frequency 6.85GHz. To realize the specified frequency allocation, the low-impedance section of MMR is estimated as $27\ \Omega$ with quarter-wavelength $\lambda_{g2}/4$ at the central frequency of 6.85GHz. Its corresponding dimensions are indicated in Figure 5.10 (a).

5.3.2 CPW Overlapped Structure

Figure 5.11 (a) Geometry and (b) equivalent circuit network of the overlapped inductive coupling structure

After allocating the first three resonance modes, the next step is to excite two poles below and above the center frequency of UWB via the tight inductive coupling element. Figure 5.11 (a) describes the geometry of the overlapped inductive coupling element with two

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different CPW feed lines. The discontinuity with enhanced coupling behaviors at reference planes P_1 and P_2 can be thoroughly investigated via the frequency-related K-inverter parameters as explained in Chapter 4. In addition to providing tight coupling, this structure terminated by two mismatching impedances Z_0 and Z_3 can generate two additional transmission poles as we want. Based on the similar theorem in Sec.5.2.2, these two poles can be characterized by the normalized K-reactance with respect to the two-side image impedance Z_1^* and Z_2^* in Figure 5.11 (b), i.e., $\bar{K} = K / \sqrt{Z_1^* Z_2^*}$.

Figure 5.12 Frequency related normalized K-reactance under different coupled distance d

Now, let us investigate this normalized K-reactance behavior over a frequency range 3.0-11.0GHz. Figure 5.12 shows four curves of this reactance versus coupled distance d under fixed coupled length $T=170\text{mil}$ around quarter-wavelength. We can observe that regardless of different distances d , the $\bar{K}$ goes up first and then falls down. The peaks occur at the frequency where two side electrical lengths are equal $\pi/2$. As d decreases,

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the coupling peak gradually rises up from 0.45 to 1.15, thus demonstrating an enhanced coupling degree. It can be further found that when d decreases to 2mil, this peak goes beyond unity, thus generating two transmission poles where $\bar{K} = 1$. So, together with the MMR three resonance frequencies discussed in previous section, a five-pole UWB filter can be realized.

5.3.3 UWB Bandpass Filter

Figure 5.13 (a) Schematic layout (b) equivalent circuit topology of proposed UWB filter

Figure 5.13 (a) is the top view of proposed UWB filter, it is composed of a single MMR and two distributed overlapped inductive coupling structures. Figure 5.13 (b) shows its equivalent circuit topology used to carry out an overall optimization design in terms of MMR and $\bar{K}$ parameters derived before.

Figure 5.14 Simulated magnitude of S_{21} under different coupled distance d

Firstly, the dimensions of MMR are determined under the very weak coupling degree, i.e., $d=15\text{mil}$, as shown in Figure 5.14. Then the target is to adjust the coupling degree in order to minimize the insertion loss within the UWB passband. Figure 5.14 depicts the frequency response of insertion loss S_{21} magnitude versus different d . As $d=15\text{mil}$, three resonant frequencies are located around 4.2, 7.0 and 9.8GHz, respectively. As d further decreases to 2 mil, the lower cut-off frequency is moving toward the targeted 3.1GHz, while the upper cut-off frequency is moving toward the targeted 10.6GHz and the magnitude $|S_{21}|$ rises to around 0-dB.

5.3.4 Experimental Verification

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Figure 5.15 Measured and Predicted results of proposed short-ended type UWB filter
(a) magnitudes of S-parameters (b) group delay

Finally, optimized UWB filter with $d=2\text{mil}$, $T=168\text{mil}$, other dimensions as shown Figure 5.10 (a) is designed based the cascaded network and its predicted electrical performances are verified by the ADS-Momentum and experiment of the fabricated circuit sample as shown in Figure 5.15. The overall length of the proposed UWB filter is 750mil. Three sets of results are further plotted together to provide a quantitative comparison. Over the

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wide frequency range 1GHz-13GHz, three sets of results are found in good agreement with each other. In the measurement, the lower and higher cutoff frequency is 3.4GHz to 10.3GHz. The maximum insertion loss and minimum return loss in UWB passband are 1.5dB and 6.5dB, respectively. Figure 5.15 (b) also shows us the measured group delay within the UWB passband varies from 0.30 to 0.65 to achieve a maximum variation 0.33ns thus implying a good phase linearity of our proposed UWB filter.

5.4 Conclusion

In this chapter, two types of novel UWB bandpass filters with enhanced interdigital CPW element and overlapped inductive coupling structure are presented. The most remarkable idea discussed is to make use of the first three multiple-mode resonances in a non-uniform CPW resonator together to build up such an ultra-wide passband. CPW interdigital or overlapped coupling structures are characterized to demonstrate the emergence of other two transmission poles. The extracted network parameters of these elements are then utilized to carry out a manual optimization of such UWB filter. By carefully allocating three resonant frequencies around 4.2 GHz, 6.85GHz and 9.8GHz within the concerned UWB band, while generating the transmission poles of CPW coupling elements around 5GHz and 9GHz, UWB passband behaviour can be realized. After predicted results are confirmed by direct fullwave simulation of its overall layout by Agilent ADS, these two types of UWB filters are fabricated. Measured results are shown in good agreement with predicted ones.

Chapter 6

Complex Per-Unit-Length Transmission Parameters of CPW Metamaterials

Abstract: Transmission line metamaterials on coplanar waveguide with series-capacitive and shunt-inductive loading in periodical intervals are thoroughly characterized using the fullwave self-calibrated method of moments, i.e., MoM-SOC technique. Much effort is made in this work to derive the two effective per-unit-length transmission parameters, i.e., complex propagation constant and characteristic impedance, aiming at demonstrating the forward- and backward-wave propagation properties in several distinctive left- and right-handed stopbands and passbands. In particular, in the whole left-handed passband, the propagation constant is exhibited purely negative phase constant while the characteristic impedance has the only positive real quantity. Thus, both of the effective

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permittivity ϵ_{eff} and permeability μ_{eff} become purely negative real in this passband. The transmission performances of the finite-cell metamaterial blocks are then characterized under the assumption of perfect impedance matching at two terminals. Finally, the two 9-cell metamaterial circuits, fed by the actual 50Ω feed lines are designed and implemented for experimental validation.

6.1 Introduction

The artificial materials or metamaterials with simultaneously negative permittivity (ϵ) and permeability (μ) or negative refractive index (n) were originally presented in theory by Veselago in 1967 [62]. Such metamaterials based on the realizable composite structures, which consist of split ring resonators and continuous wires, were firstly introduced by Pendry et al. [63]. One of their promising applications in optics is to explore the perfect optical lens [120, 121]. In 2001 [64], these metamaterials were successfully implemented to experimentally confirm the existence of negative effective refractive index in a certain frequency passband. Since then, it has been dramatically arousing a great interest in the optical and electromagnetic societies. In addition to the above bulky three-dimensional (3-D) metamaterial structures, the 1-D and 2-D transmission line based metamaterials were constituted in [60, 61] by periodically interrupting the uniform microstrip line with the capacitor and inductor chips in series and shunt formats. In this aspect, the dispersion diagram, i.e., phase constant versus frequency in the passband, of a periodic transmission line based on microstrip line and coplanar waveguide has been derived to demonstrate some of basic guided-wave

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propagation properties in the passbands, such as negative phase constant or anti-parallel phase and group velocities, as investigated in [60-61, 122-125] using the classical periodic transmission line theorem [54]. By the way, the similar periodic transmission line topology was studied several decades ago to generate the called “backward waves” with the opposite signs in phase and group velocities, e.g. [127]. The recent works [126, 128] have given the more detailed description on the characterization of these ideal or non-dispersive transmission line metamaterials. On the other hand, extensive works have been carried out to explore various metamaterials-based planar circuits and antennas with advanced performances and miniaturized dimension, such as dual-band branch-line couplers, zeroth-order transmission line resonators, and backward leaky-wave antennas, as summarized in [66].

Nevertheless, to our knowledge, little effort has been reportedly made today to characterize the fully integrated and inhomogeneous transmission line metamaterials in terms of effective permittivity ϵ_{eff} and permeability μ_{eff} or effective per-unit-length series capacitance and shunt inductance of the 1-D composite metamaterials with infinite extent. Figure 6.1(a) and 6.1 (b) show the geometrical configurations of the periodic unit cell of the two integrated CPW metamaterials without installing any external capacitor/inductor chips. The main objective of this work is to derive the two effective per-unit-length transmission parameters, i.e., characteristic impedance and propagation constant of these integrated CPW metamaterials, thus originally demonstrating their fundamental forward- and backward-wave propagation or attenuation properties in the left-/right-handed passbands and stopbands.

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To accomplish this task, the hybrid fullwave approach based on the method of moments (MoM) and short-open calibration (SOC) procedure described in Chapter 2 and [52, 129-132], namely, MoM-SOC technique is employed to numerically de-embed these per-unit-length parameters with dispersive and/or complex quantities in a wide frequency range. So far, this MoM-SOC has achieved a great success in numerically de-embedding the S-parameters of the device under test (DUT) [129, 131] and extracting the equivalent circuit model parameters of various planar circuit elements as well as the per-unit-length transmission parameters of uniform CPW and CPW electromagnetic bandgap (EBG) structures [59, 130]. The remainder of this paper is organized as follows. In Section II, the equivalent π - and T-networks of CPW discontinuities and per-unit-length transmission parameters of the uniform CPW are extracted by executing the full-wave MoM-SOC approach to establish a simple cascaded network theory for characterization of the 1-D CPW metamaterials with infinite extent. In Section III, the CPW metamaterials with finite extent, terminated by the perfect impedance matching feed lines, are characterized to demonstrate the guided-wave propagation across the finite-extended metamaterial region without receiving the multiple reflections at its two terminals. In final, the two finite-extent CPW metamaterials with the realizable 50- Ω feed lines are modelled and the two CPW circuits are accordingly fabricated to provide an experimental verification on the predicted wave transmission behaviours under the multiple reflections.

6.2 Infinite-Extent CPW Metamaterials

6.2.1 Geometry and Equivalent Topology

Figure 6.1 The two types of coplanar waveguide metamaterial structures to be characterized. (Gray area is conductor, white area is slot, $t_c=t_i=0.127\text{mm}$, $T=8.382\text{mm}$)
 (a), (c) layout and equivalent circuit model of unit cell of the type-A structure; (b) and (d) layout and equivalent circuit model of unit cell of the type-B structure.

Figure 6.1(a) and 6.1(b) depict the layouts of the two types of fully integrated CPW metamaterials with the repeated unit cells of periodicity (T). Figure 6.1(c) and 6.1(d) indicate the two relevant equivalent transmission line networks of the two periodical cells. To show the basic topology in conjunction with the called 1-D composite transmission line metamaterials with simultaneously series-capacitive and shunt-inductive loading [65-

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67, 123, 128], only the major shunt inductance of CPW shorted-stub and the major series capacitance of the CPW opened gap are shown here. As discussed in Chapter 3, these two CPW lumped-element discontinuities can be in general characterized as the two complete equivalent circuit networks. In this aspect, the CPW opened-gap is considered as the π -type network with one series susceptance, $\bar{B}_g$, at center and two shunt susceptance, $\bar{B}_p$, at two sides. Meanwhile, the CPW shorted-stub is perceived as the T-type network with one shunt reactance, $\bar{X}_g$, at center and two series reactance, $\bar{X}_p$, at two sides. Using the MoM-SOC technique, all the above network parameters can be numerically extracted as executed in Sec. 4.2. Moreover, the dispersive phase constant and characteristic impedance of the uniform CPW can be derived in this MoM-SOC framework as done in Sec.3.5.

In order to simplify the analysis on a basis of the two simple topologies in Figure 6.1(c) and 6.1(d), the above-discussed π - or T-type network is further expressed as the circuit model with the shunt reactance (X) or series susceptance (B) at center and the two equal electrical lengths, ϕ_l or ϕ_c , at two sides. Under the equivalence of two related networks for the same CPW circuit, the normalized reactance ($\bar{X}$) and the normalized susceptance ($\bar{B}$) as well as ϕ_l and ϕ_c can be analytically derived.

In order to obtain the transformation formulae, we can choose the excitations so that the central symmetry plane corresponds to an electrical wall (short circuit) or magnetic wall (open circuit). At the central symmetry plane, half structure of π network consists of one shunt susceptance B_p , and a series susceptance $2B_g$, while the network in Figure 6.1 (c)

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consists of one series susceptance $2B$ in the right of electrical length ϕ_c .

By equating the two network characteristic admittance at open and short circuits, the two unknown parameters B and ϕ_c can be derived:

$$Y_{in} = j(B_p + 2B_g) = Y_{ou} \frac{j2B + jY_{ou} \tan(\phi/2)}{Y_{ou} - 2B \tan(\phi/2)} \quad \text{For short circuit} \quad (6.1a)$$

$$Y_{in} = jB_p = jY_{ou} \tan \phi/2 \quad \text{For open circuit} \quad (6.1b)$$

thus:

$$\bar{B} = \frac{\bar{B}_g}{1 + 2\bar{B}_p \bar{B}_g + \bar{B}_p^2} \quad (6.1c)$$

$$\phi_c = \tan^{-1} \bar{B}_p \quad (6.1d)$$

where $\bar{B} = B/Y_{ou}$. The symbol, Y_{ou} , is the characteristic admittance of the uniform CPW transmission line.

Following the similar procedure, by equating characteristic impedance of half structure of the T-type network, composed of series reactance X_p and shunt reactance $2X_g$, and the shunt reactance network in Figure 6.1 (d) at open and short circuit case, the corresponding reactance X and electrical length ϕ_l can be derived.

$$Z_{in} = j(X_p + 2X_g) = Z_{ou} \frac{j2X + jZ_{ou} \tan(\phi/2)}{Z_{ou} - 2X \tan(\phi/2)} \quad \text{For open circuit} \quad (6.2a)$$

$$Z_{in} = jZ_p = jZ_{ou} \tan \phi/2 \quad \text{For short circuit} \quad (6.2b)$$

Thus:

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$$\bar{X} = \frac{\bar{X}_g}{1 + 2\bar{X}_p\bar{X}_g + \bar{X}_p^2} \quad (6.2c)$$

$$\phi_l = \tan^{-1} \bar{X}_p \quad (6.2d)$$

where $\bar{X} = X / Z_{0u}$, the symbol, Z_{0u} , is the characteristic impedance of the uniform CPW transmission line.

Due to the periodical configuration in essence, the considered CPW metamaterials with infinite extent can be comprehensively studied in terms of the effective propagation constant and characteristic impedance [54, 59] or effective per-unit-length series susceptance and shunt reactance [127, 128] of equivalent 1-D uniform transmission line media. To simplify the analysis, the single CPW periodic unit cells of the periodicity (T) in Figure 6.1(a) and 6.1(b) are modelled as the resultant ABCD matrices, $[A]_a$ and $[A]_b$, via matrix multiplication of all the cascaded networks as shown in Figure 6.1(c) and 6.1(d). Here, T is selected larger than $\lambda/10$ in order to minimize the unexpected mutual coupling between two adjacent CPW elements. Meanwhile, the ABCD matrix of such an effective uniform CPW transmission line with the finite length (T) can be in general expressed,

$$[A]_T = \begin{bmatrix} \cosh(rT) & Z_c \sinh(rT) \\ Y_c \sinh(rT) & \cosh(rT) \end{bmatrix} \quad (6.3)$$

6.2.2 Per-Unit-Length Transmission Parameters

Under the condition that $[A]_a = [A]_T$ and $[A]_b = [A]_T$, the complex propagation constant (γ) and complex characteristic impedance (Z_c) or admittance (Y_c) of the two types of 1-D CPW metamaterials can be explicitly obtained in terms of the parameters of the two networks in Figure 6.1(c) and 6.1(d). Regardless of either Type-A or Type-B, the propagation constants (γ) in these two structures are found exactly identical [54] with each other and they can be expressed in the same closed-form formulae.

$$\gamma T = \cos^{-1} \left[\cos(2\phi) + \frac{\sin(2\phi)}{2\bar{B}} + \frac{\sin(2\phi)}{2\bar{X}} - \frac{\cos^2(\phi)}{2\bar{B}\bar{X}} \right] + n\pi \quad (6.4)$$

where n is the integer to be selected. Equation (6.4) implies that the frequency diagrams of effective guided-wave phase and attenuation constants become basically unchanged if the periodic topology of a CPW metamaterial is constituted. But, it is not true for the effective characteristic impedance or Bloch-wave impedance [54]. The two distinctive closed-form formulas are separately given as below in conjunction with the Type-A and Type-B CPW metamaterials that are only different in the selection of two interfaces of a whole periodic unit cell.

$$Z_c = \frac{-j \sinh(\gamma T) \cdot Z_{0u}}{\sin(2\phi) - \cos^2(\phi) / \bar{X}} \quad (\text{Type-A}) \quad (6.5a)$$

and

$$Y_c = \frac{-j \sinh(\gamma T) \cdot Y_{0u}}{\sin(2\phi) - \cos^2(\phi) / \bar{B}} \quad (\text{Type-B}) \quad (6.5b)$$

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where the total electrical length (ϕ) consists of the three parts, that is, $\phi = \phi_c + \phi_t + \phi_l$, in Figure 6.1(c) and 6.1(d). $\phi_t = 2\pi l / \lambda_g$ corresponds to the CPW line with the length (l) that connects the two adjacent CPW inductive and capacitive elements. λ_g and Z_{0u} or Y_{0u} are the guided wavelength and characteristic impedance or admittance of the above CPW line, respectively.

In this work, the two CPW metamaterials, as shown in Figure 6.1(a) and 6.1(b), are formed on the substrate of RT/Duroid 6010 with the permittivity of $\epsilon_r=10.8$ and thickness of $h=0.635\text{mm}$. With the use of Equations (6.4) and (6.5), the effective complex propagation constant ($\gamma=\alpha+j\beta$) and characteristic impedance ($Z_c=\text{Re}(Z_c)+j\text{Im}(Z_c)=R_c+jX_c$) are numerically obtained. Figure 6.2(a) depicts the normalized propagation constant (γ/β_0) of both the two types of CPW metamaterials, where β_0 is the free space wave number. Figure 6.2(b) and 6.2(c) are the two different effective characteristic impedances (Z_c) of these two CPW metamaterials. The results in Figure 6.2(a) to 6.2(c) have originally demonstrated the most fundamental guided-wave propagation properties of the fully integrated CPW metamaterials in terms of dispersive and complex per-unit-length parameters, i.e., γ and Z_c . As illustrated in Figure 6.2, there exist the four distinct frequency regions with stopband and passband characteristics [65, 66]. According to the signs of the phase constant (β), they can be classified into the left-handed (LH) stopband ($\beta<0$, $\alpha>0$), LH passband ($\beta<0$, $\alpha=0$), stopband ($\beta=0$, $\alpha>0$) and right-handed (RH) passband ($\beta>0$, $\alpha=0$). In Figure 6.2, the two distinct passbands are filled by the grey color to distinguish the passbands and stopbands.

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(a)

(b)

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(c)

Figure 6.2 Frequency dispersions of extracted effective per-unit-length transmission parameters of the two CPW metamaterials. (CPW strip width: 1.016mm; CPW slot width: 0.889mm) (a) Propagation constant γ/β_0 (Type-A & Type-B); (b) (c) Characteristic impedance Z_c (type-A & type-B).

Now, let's focus on the investigation on the guided-wave properties with reference to the results in Figure 6.2(a) to Figure 6.2(c). In the LH stopband, $\alpha > 0$ and $\beta T = -\pi$ are valid, indicating that the guided-wave attenuates as the “backward-wave” [127, 128]. As the frequency increases in this stopband, the normalized α/β_0 rapidly decreases from the large positive value to zero while β/β_0 rises up linearly in the negative region. Due to $\beta T = -\pi$, $\sinh(\gamma T) = -\sinh(\alpha T) < 0$. Thus, the impedance (Z_c) in type-A and admittance (Y_c) in Type-B, in Equations (6.5a) and (6.5b), become purely imaginary, i.e.,

$$Z_c = \frac{j \sinh(\alpha T) \cdot Z_{0u}}{\sin(2\phi) - \cos^2(\phi) / \bar{X}} \quad (\text{Type-A}) \quad (6.6a)$$

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$$Y_c = \frac{j \sinh(\alpha T) \cdot Y_{0u}}{\sin(2\phi) - \cos^2(\phi) / \bar{B}} \quad (\text{Type-B}) \quad (6.6b)$$

As can be seen in Figure 6.2(b) and 6.2(c), Z_c of Type-A tends to linearly increase from the negative to zero with pure imaginary value, while the reactance of Type-B quickly increases from the finite positive to the infinity as the frequency is raised to 4.9GHz that is the lower edge of the LH passband.

In the LH passband, the negative β/β_0 consistently goes up and reaches to zero with frequency while $\alpha/\beta_0=0$. Thus, it implies the basic performance of the “backward-wave” with opposite or anti-parallel group velocity ($v_g = d\omega / d\beta$) and phase velocity ($v_p = \omega / \beta$), as discussed in [124-128]. Due to $\alpha=0$ in this band, $\sinh(\gamma T) = j \sin \beta T$.

Thus, Equations (6.5a) and (6.5b) can be simply expressed in terms of the negative β .

$$Z_c = \frac{\sin(\beta T) \cdot Z_{0u}}{\sin(2\phi) - \cos^2(\phi) / \bar{X}} \quad (\text{Type-A}) \quad (6.7a)$$

$$Y_c = \frac{\sin(\beta T) \cdot Y_{0u}}{\sin(2\phi) - \cos^2(\phi) / \bar{B}} \quad (\text{Type-B}) \quad (6.7b)$$

From Equations (6.7a) and (6.7b), Z_c in both CPW metamaterials have the purely positive real value or resistance as can be confirmed in Figure 6.2(b) and 6.2(c). This resistance starts to go up from zero and then falls down to zero across the single peak in the Type-A metamaterial. Meanwhile, it monotonously drops off from the positive infinity to zero in the Type-B metamaterial. As the frequency further increases, the guided wave moves to the middle stopband that covers the frequency range from 5.8 to 7.2GHz. In this stopband,

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because of $\beta T=0$ instead of $\beta T=-\pi$ in the LH stopband, the negative “-” sign needs to be added into Equations (6.6a) and (6.6b) so as to calculate the impedance (Z_c) with purely non-zero reactances. This reactance appears to rise up towards the infinity in both cases. As the frequency is larger than 7.2GHz, the guided-wave enter into the second passband with $\beta>0$ and $\alpha=0$, indicating the propagation property of the “forward-wave” [127] or right-handed wave with parallel phase and group velocities. In this band, the impedances in both cases become the purely positive real values again with the frequency dispersions shown in Figure 6.2(b) and 6.2(c). Beyond this RH passband, RH stopbands and passbands with increased positive β successively come forth as discussed in many early literatures, e.g., [54], for periodic structures.

6.2.3 Effective Permittivity & Permeability of CPW Metamaterials

To provide a comprehensive insight into the above two CPW metamaterials, it is definitely imperative for one to derive the effective per-unit-length shunt capacitance C_{eff} and series inductance L_{eff} of a periodic transmission line or effective permittivity ϵ_{eff} and permeability μ_{eff} of a 1-D composite material. On a basis of the simple transmission line and uniform plane-wave propagation theorems [54, 125], the propagation constant γ and characteristic impedance Z_c discussed above can be straightforwardly expressed in terms of either C_{eff} and L_{eff} or ϵ_{eff} and μ_{eff} such that,

$$\gamma = j\omega\sqrt{L_{eff}C_{eff}} = j\omega\sqrt{\mu_{eff}\epsilon_{eff}} \quad (6.8a)$$

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$$Z_c = \sqrt{\frac{L_{eff}}{C_{eff}}} = \sqrt{\frac{\mu_{eff}}{\epsilon_{eff}}} \quad (6.8b)$$

After a simple analysis is undertaken to Equations (6.7a) and (6.7b), the two complex parameters, $\epsilon_{eff} = \text{Re}[\epsilon_{eff}] + j \text{Im}[\epsilon_{eff}]$ and $\mu_{eff} = \text{Re}[\mu_{eff}] + j \text{Im}[\mu_{eff}]$, can be expressed in terms of real and imaginary parts of complex characteristic impedance $Z_c = R_c + jX_c$ and propagation constant $\gamma = \alpha + j\beta$.

$$\mu_{eff} = \frac{(\beta R_c + \alpha X_c)}{\omega} - j \frac{(\alpha R_c - \beta X_c)}{\omega} \quad (6.9a)$$

$$\epsilon_{eff} = \frac{(\beta R_c - \alpha X_c)}{\omega(X_c^2 + R_c^2)} - j \frac{(\alpha R_c + \beta X_c)}{\omega(X_c^2 + R_c^2)} \quad (6.9b)$$

By substituting the derived γ and Z_c (Figure 6.2) to Equations (6.8a) and (6.8b), μ_{eff} and ϵ_{eff} of the two considered CPW metamaterials with infinite extent can be simply calculated and they are plotted in Figure 6.3 and Figure 6.4 for the Type-A and Type-B cases, respectively.

In the LH stopband below 4.9GHz, Equations (6.9a) and (6.9b) can be simplified as below due to $R_c = 0$.

$$\mu_{eff} = \frac{\alpha X_c}{\omega} + j \frac{\beta X_c}{\omega} \quad (\text{LH stopband}) \quad (6.10a)$$

$$\epsilon_{eff} = -\frac{\alpha}{X_c \omega} - j \frac{\beta}{\omega X_c} \quad (\text{LH stopband}) \quad (6.10b)$$

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Equations (6.10a) and (6.10b) imply that both real and imaginary parts of μ_{eff} and ε_{eff} become non-zero in this lowest stopband regardless of either CPW metamaterials. Moreover, considering that $\alpha > 0$, $\beta < 0$ and $X_c < 0$ (Type-A) or $X_c > 0$ (Type-B) shown in Figure 6.2, μ_{eff} and ε_{eff} have the opposite polarities as can be observed in Figure 6.3 and Figure 6.4.

In the LH passband of 4.9 to 5.8GHz, we can find from Figure 6.2 that $\alpha = 0$ and $X_c = 0$ are definitely valid. As such, Equations (6.9a) and (6.9b) become,

$$\mu_{eff} = \frac{\beta R_c}{\omega} \quad (\text{LH/RH passband}) \quad (6.11a)$$

$$\varepsilon_{eff} = \frac{\beta}{\omega R_c} \quad (\text{LH/RH passband}) \quad (6.11b)$$

Equations (6.11a) and (6.11b) indicate that both ε_{eff} and μ_{eff} have the purely real quantities in the LH passband. Plus that $\beta < 0$ and $R_c > 0$ as shown in Figure 6.2 in both cases, the real quantities of ε_{eff} and μ_{eff} become simultaneously negative with zero imaginary quantities, thereby exhibiting the negative index of refraction (n) [64] of these two CPW metamaterials.

$$n = -\sqrt{\frac{\mu_{eff}\varepsilon_{eff}}{\mu_0\varepsilon_0}} < 0 \quad (\text{LH passband}) \quad (6.12)$$

where ε_0 and μ_0 are the permittivity and permeability in air.

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Figure 6.3 Frequency dispersion of (a) effective complex permeability (μ_{eff}) and (b) effective complex permittivity (ϵ_{eff}) of the CPW metamaterial: Type-A

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Figure 6.4 Frequency dispersion of (a) effective complex permeability (μ_{eff}) and (b) effective complex permittivity (ϵ_{eff}) of the CPW metamaterial: Type-B

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In the middle stopband of 5.8 to 7.2GHz, $\beta = 0$ plus $R_c = 0$ that is valid in the LH stopband. As such, Equations (6.10a) and (6.10b) can be further simplified with only real quantities, i.e.,

$$\mu_{eff} = \frac{\alpha X_c}{\omega} \quad (\text{middle stopband}) \quad (6.13a)$$

$$\varepsilon_{eff} = -\frac{\alpha}{X_c \omega} \quad (\text{middle stopband}) \quad (6.13b)$$

Equations (6.13a) and (6.13b) imply that the imaginary parts of both μ_{eff} and ε_{eff} wholly disappear in the middle stopband as also illustrated in Figure 6.3 and Figure 6.4. In particular, $\text{Re}[\mu_{eff}]$ and $\text{Re}[\varepsilon_{eff}]$ always have the opposite polarities in both CPW structures, thus bringing out the wave attenuation with $\alpha > 0$ that is shown in Figure 6.2(a).

In the RH passband, μ_{eff} and ε_{eff} can still be derived based on Equations (6.11a) and (6.11b). However, the conditions of $\beta > 0$ and $R_c > 0$ lead to simultaneously positive $\text{Re}[\mu_{eff}]$ and $\text{Re}[\varepsilon_{eff}]$ while $\text{Im}[\mu_{eff}] = 0$ and $\text{Im}[\varepsilon_{eff}] = 0$ as shown in Figure 6.3 and Figure 6.4 for the two considered CPW metamaterials.

6.3 Finite-Extent CPW Metamaterials

6.3.1 Truncated Transmission Line Model

In this section, the above infinite-extent CPW metamaterials are truncated longitudinally

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and fed by the two CPW feed lines with equivalent source and load impedances, Z_s and Z_L . The magnitude and phase of transmission coefficient, i.e., $|S_{21}|$ and ϕ_{21} , of these finite-extent CPW metamaterials are studied using the simple cascaded transmission line theorem [54, 125]. The objective here is to demonstrate the propagation properties of the forward- and backward-wave traveling across these finite-extent metamaterials without and with receiving the multiple reflections at the two truncated interfaces. Figure 6.5(a) depicts the first case where the source and load impedances are assumed to be the exactly same as the dispersive and complex characteristic impedance of 1-D transmission line metamaterials, i.e., $Z_s=Z_L=Z_c$. Regardless of either stopband or passband, the transmission coefficient (S_{21}) of this truncated structure can be simply expressed in terms of the two above-derived complex propagation constant, $\gamma=\alpha+j\beta$, for the N unit cells or the finite length of NT.

$$S_{21} = e^{-(\alpha+j\beta)NT} \quad (6.14)$$

Equation (6.14) can be further simplified such that $|S_{21}|$ and ϕ_{21} can be expressed in the two separate closed-form formulae.

$$|S_{21}| = e^{-\alpha NT} \quad (6.15a)$$

$$\phi_{21} = -\beta NT \quad (6.15b)$$

In the actual implementation, the CPW feed lines with dispersive and complex characteristic impedances can not be really constructed. For this reason, let's consider the truncated CPW metamaterials connected to the two uniform CPW lines with the almost

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Figure 6.5 Diagrams of guided-wave propagation in the finite-extent transmission line metamaterials with the feed lines of: (a) $Z_s = Z_L = Z_c$; (b) $Z_s \neq Z_c$ & $Z_L \neq Z_c$.

unchanged source and load impedances, i.e., $Z_s = Z_L \neq Z_c$. As illustrated in Figure 6.5(b), the multiple reflections may happen at the two truncated interfaces such that the guided-wave in the metamaterial region is bounced ceaselessly. So, the resultant transmission coefficient (S_{21}) of these finite-extent CPW metamaterials can be obtained based on the cascaded transmission line theorem [54, 125] such that

$$S_{21} = \frac{T_1 T_2 e^{-\gamma NT}}{1 + \Gamma_1 \Gamma_2 e^{-2\gamma NT}} \quad (6.16)$$

where Γ_1 and T_1 , Γ_2 and T_2 indicate the two sets of transmission and reflection coefficients at the two interfaces, #I and #II, along the source-to-load direction, respectively, and they can be expressed in terms of the three impedances, Z_s , Z_c and Z_L .

$$T_1 = \frac{2Z_c}{Z_s + Z_c} \text{ and } \Gamma_1 = \frac{Z_c - Z_s}{Z_c + Z_s} \quad (6.17a)$$

$$T_2 = \frac{2Z_L}{Z_L + Z_c} \text{ and } \Gamma_2 = \frac{Z_L - Z_c}{Z_L + Z_c} \quad (6.17b)$$

6.3.2 Predicted Transmission Performances

Using Equations (6.14) and (6.16), the magnitudes ($|S_{21}|$) and phases (ϕ_{21}) of the two finite-extent CPW metamaterials with the unit-cell configurations, shown in Figure 6.1(a) and 6.1(b), can be separately calculated under the impedance matching and impedance-mismatching situations as discussed above. Figure 6.6(a) and 6.6(b) depict the three sets of these two parameters versus frequency under $N=9$. In the perfectly matching case with $Z_s=Z_L=Z_c$, the guided-wave in the LH and RH passbands achieves the full transmission with $|S_{21}|=0\text{dB}$ and opposite polarities of ϕ_{21} while the phases ϕ_{21} in the LH and middle stopbands become the constant values, i.e., 9π and 0. In particular, we can see from Figure 6.6(b) that the positive ϕ_{21} in the LH passband is gradually reduced from 9π (LH stopband) to 0 (middle stopband) while the negative ϕ_{21} in the RH passband decreases from 0 (middle stopband) to -9π (1st RH stopband beyond 8.0GHz).

Now, let's look at the two practical cases in which the two CPW finite-extent CPW metamaterials are fed by the two 50- Ω feed lines. Due to $Z_s=Z_L \neq Z_c$, the magnitude ($|S_{21}|$) and phase (ϕ_{21}) are not only predominantly determined by the complex propagation constant (γ) but also are seriously affected by the unequal impedances (Z_s , Z_L , Z_c) as well as the finite length (NT). Moreover, due to the difference in effective

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Figure 6.6 Simulated transmission coefficients of the three finite-extent CPW metamaterials with $N=9$. (a) magnitude ($|S_{21}|$), (b) phase (ϕ_{21}).

characteristic impedance between the type-A and type-B, $|S_{21}|$ and ϕ_{21} become slightly different as illustrated in Figure 6.6(a) and 6.6(b), in which they are separately indicated with the dashed and solid lines. Generally speaking, all the three sets of graphs, in Figure

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6.6(a) and 6.6(b), are very similar in configuration with each other in the plotted stopbands and passbands. In the two stopbands, ϕ_{21} in Type-A and Type-B cases are slightly shifted below 9π and 0 in the perfectly matching case. It can be interpreted that the non-zero imaginary parts of Z_c bring out an additional frequency-related non-zero phase at the two truncated interfaces. On the other hand, several ripples can be clearly observed in both LH and RH passbands. In the stopbands, the guided-wave appears to attenuate as an exponential function of $(-\alpha NT)$: $N=9$ in this case). It implies that such an evanescent guided-wave incident at one interface may almost disappear at the other interface for the lengthened finite length (NT). But, it is not correct in the two passbands. As explained above via Equation (6.16), the guided wave in the LH and RH passbands are ceaselessly bounced between the two interfaces, #1 and #2, of the truncated metamaterial section, thus creating the multiple ripples in $|S_{21}|$ and ϕ_{21} .

6.3.3 Experimental Verification

In final, the two finite-extent CPW metamaterials with the two 50Ω CPW feed lines are characterized and fabricated to provide an experimental verification on the predicted performances. Figure 6.7(a) and Figure 6.8 (a) depict the layouts of the two constructed metamaterial-based two-port circuit blocks. Herein, the additional short CPW line sections are introduced between the CPW 50Ω feed lines and metamaterial regions at each truncated interfaces. Figure 6.7(b) and Figure 6.8(b) plot the magnitudes and phases of the predicted and measured transmission coefficients over the frequency ranges of

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Figure 6.7 Predicted and measured transmission coefficients of the finite-extent CPW metamaterials (Type-A). (a) Layout; (b) Results.

covering the LH stopband, LH passband, middle bandstop and RH passband. The measured parameters are de-embedded at the two specified reference planes that are denoted by the dotted lines in Figure 6.7(a) and 6.8(a). Both predicted and measured results are in reasonable agreement with each other, the losses in experiment are mainly caused by the substrate loss, conductor loss, radiation loss and unexpected mismatching effect at the SMA connector. All of them exhibit the distinct transmission characteristics of the LH and RH passbands. Of the most importance, the positive ϕ_{21} in the LH passband and negative ϕ_{21} in the RH passband demonstrate us that the guided-wave

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propagation in these two passbands is undertaken in terms of “backward-wave” and “forward-wave” [54, 125], respectively.

Figure 6.8 Predicted and measured transmission coefficients of the finite-extent CPW metamaterials (Type-B). (a) Layout; (b) Results.

6.4 Conclusion

In this chapter, the 1-D CPW transmission line metamaterials are investigated to demonstrate the forward- and backward-wave propagation properties in the two distinctive passbands. Using the self-calibrated method of moments, namely, MoM-SOC

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approach, the two effective per-unit-length transmission parameters, i.e., propagation constant and characteristic impedance, are numerically extracted. Thus, it allows one to explicitly derive the effective permittivity (ϵ_{eff}) and permeability (μ_{eff}) of the 1-D composite metamaterials. Extensive results on these periodic transmission line parameters and 1-D metamaterial parameters are provided over a wide frequency range to originally demonstrate the fundamental guided-wave propagation and attenuation characteristics in the LH stopband, LH passband, middle stopband and RH passband. In particular, it is exhibited that in the LH passband, ϵ_{eff} and μ_{eff} become simultaneously negative pure real quantities, thus showcasing the negative index of refraction (n). In the meantime, the opposite and same polarities of the phase constant (β) indicate that the guide-wave propagates in the LH and RH passbands in the two different ways of the “backward-wave” and “forward-wave”. Moreover, the finite-extent CPW metamaterials are studied on a basis of the simple cascaded transmission line theorem. The magnitudes and phases of transmission coefficient (S_{21}) of 9-cell finite-extent CPW metamaterials with and without multiple reflections at the two truncated interfaces are characterized. In final, the two CPW circuit blocks with 50- Ω feed lines are designed and fabricated to provide an experimental verification on the predicted performances of the two considered CPW metamaterials.

Chapter 7

CPW Metamaterials with Varied LH and RH Passbands

Abstract: Coplanar waveguide (CPW) metamaterials with varied left- (LH) and right-handed (RH) passbands are extensively studied in terms of per-unit-length transmission parameters and magnitude/phase of transmission coefficient, respectively. The two CPW metamaterials are constructed by simultaneously loading the lumped or distributed CPW series-capacitive and shunt-inductive elements in periodic intervals. Effective characteristic impedance and propagation constant of these infinite-extent metamaterials are first extracted to give a physical insight into the guided-wave propagation characteristics in the LH/RH passbands. It is exhibited that the stopband between the LH and RH passbands can be wholly eliminated in the distributed CPW case. The two finite-extent CPW metamaterials with 50Ω feed lines are then modeled via simple cascaded transmission line theorem. Predicted magnitude and phase of transmission coefficient are

confirmed by experiment, showing the distinctive backward- and forward-wave performances in the LH and RH passbands.

7.1 Introduction

As introduced in Chapter 6, metamaterials with simultaneously negative permittivity (ϵ) and permeability (μ) originally predicted in theory by Veselago have attracted more and more research interests [62]. Since the three-dimensional (3-D) bulky metamaterial has been innovatively constructed to provide the first experimental verification on the existence of negative effective refractive index in a certain frequency range [64], a tremendous attention has been aroused towards proposal, modelling and implementation of various 2-D and 1-D planar microstrip line metamaterials [121, 128] for applications in microwave engineering, as summarized in [66]. Also, the 1-D metamaterials can be formulated on coplanar waveguide (CPW) by periodically installing the capacitor and inductor chips [121]. Nevertheless of different configurations, all the 1-D transmission line metamaterials can actually be referred as to the “backward wave” infinite-extent transmission line with the anti-parallel phase and group velocities [133]. And further, several numerical methods are proposed in modelling different planar metamaterial structures recently [134-137], however, to my knowledge, little literature has thoroughly investigated on how to determine the varied left-handed (LH) passband and right-handed (RH) passband for newly potential application, such as UWB filter and scanning antenna, on the fully integrated metamaterials without adding additional chips.

In this chapter, two lumped and distributed capacitive and inductive coupling elements

are firstly analyzed via the 1-D metamaterial series capacitor and shunt inductor network, respectively, as shown in Figure 7.1. Based on the analysis methodology discussed in Chapter 6, closed-form formulae for the cut off frequency of LH/RH passbands are obtained through the frequency dispersive relationship of guided-wave propagation constant. Next, the 1-D transmission line metamaterials on CPW are studied to demonstrate the fundamental guided-wave propagation characteristics in the infinite-extent 1-D media in terms of effective per-unit-length characteristic impedance or Bloch-wave impedance [137] and propagation constant [54] via MoM under different lumped/distributed element dimensions. Finally, using the simple cascaded transmission line network, the magnitudes and phases of transmission coefficients of the two finite-extent 1-D CPW metamaterials with nine unit cells are further derived to provide an experimental verification on the predicted performances. Of importance, the results quantitatively show that the bandstop between the distinctive LH and RH passbands can be fully eliminated by utilizing the distributed CPW series-capacitive and shunt-inductive elements with enhanced quantities. It may be required in many applications, e.g., steerable backfire-to-endfire leaky wave antennas [66].

7.2 CPW Discontinuities involved in 1-D Metamaterials

7.2.1 Geometry of CPW Elements

Figure 7.1 (a) and (b) depict the geometry of lumped capacitive and inductive CPW element as two basic components of 1-D CPW metamaterials unit cell. Figure 7.1 (c) and

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(d) describe the distributed CPW elements with enhanced coupling. As discussed in Sec. 6.2.1, the lumped and distributed capacitive coupling element in Figure 7.1 (a) (c) can be modelled as the series susceptance (B) at center and the two equal electrical lengths, ϕ_c at two sides as Figure 7.1(e). Meanwhile the inductive coupling elements in Figure 7.1 (b)

Figure 7.1 Geometries and equivalent models of the capacitive and inductive coupling CPW discontinuities on the substrate of $\epsilon_r=10.8$ and $h=0.635\text{mm}$. (a) (b) Lumped CPW elements; (c),(d) Distributed CPW elements (e) and (f) equivalent series capacitor and shunt inductor transmission line circuit model of 1-D metamaterial media

and (d) can be in general modelled as shunt reactance (X) in the center, and two equal electrical lengths ϕ_l at two sides as Figure 7.1 (f).

7.2.2 Series Susceptance and Shunt Reactance

(a)

(b)

Figure 7.2 Conventional capacitive coupling element network parameters (a) normalized series susceptance B/Y_{0u} , (b) equivalent electrical length ϕ_c

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Figure 7.3 Distributed capacitive coupling element network parameters (a) normalized series susceptance B/Y_{0u} , (b) equivalent electrical length ϕ_c

Based on the work in Sec. 6.2.1, the normalized series susceptance $\bar{B}$ and electric length ϕ_c can be obtained via π type network parameters of capacitive coupling elements. Figure 7.2 depicts these two parameters frequency dispersive curves under three different coupled gap G_1 with slot width of 15mil and central strip width of 50mil feed lines. The

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normalized susceptance $\bar{B}$ is observed as a linear function of frequency to show a frequency independent series capacitor in the denoted frequency range. As the coupled gap G_1 decreases, the equivalent series capacitor increases while the edge electric length ϕ_c becomes shorter. In some cases, such capacitor requires to be further increased, so that a distributed capacitive coupling element shown in Figure 7.1 (c) is utilized to provide enhanced capacitor value. The normalized susceptance $\bar{B}$ of distributed capacitive coupling element in Figure 7.3 (a) exhibits much larger value than that of the conventional coupling element as shown in Figure 7.2 (a). As the length l_2 extended, the normalized $\bar{B}$ goes up at certain frequency with fixed gap width G_3 , while the phase ϕ_c exhibits as large as around ten times of the value in the conventional lumped coupling element and it also increases with extended l_2 .

Following the formulae in (6.2c) and (6.2d), the lumped and distributed inductive coupling elements are expressed as shunt reactance X and two equal phases ϕ_l . Figure 7.4 plots the normalized reactance $\bar{X}$ and phase ϕ_l as a function of frequency under three different gap widths G_2 . The linear frequency response demonstrates that the frequency independent shunt inductance increases with the shortened G_2 . Figure 7.5 depicts the normalized reactance $\bar{X}$ and ϕ_l curves under three different stub length d_2 . It can be indicated, that at smaller length d_2 , the reactance $\bar{X}$ increases linearly with frequency. As d_2 further increases to 60mil, this reactance increases gradually at first to its maximum value and then falls down at higher frequency. Meanwhile phase ϕ_l increases as a linear function of frequency in the considered frequency range (4-10GHz).

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(a)

(b)

Figure 7.4 Lumped inductive coupling element network parameters (a) normalized series reactance X/Z_{0u} , (b) equivalent electrical length ϕ_l

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(a)

(b)

Figure 7.5 Distributed inductive coupling element network parameters (a)

normalized series reactance X/Z_{0u} , (b) equivalent electrical length ϕ_l

7.3 Per-unit-Length Parameters of CPW Metamaterials

In the previous section, the lumped/distributed capacitive elements have been characterized as the circuit models that with one series reactance (B) in the center and the

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two same electrical lengths (ϕ_c) at either side as indicated in Figure 7.1 (e). Similarly, the CPW inductive elements can be modelled as the shunt susceptance (X) and the two same electrical lengths (ϕ_l) as shown in Figure 7.1 (f). Thus, the unit cell of metamaterial can be expressed as the equivalent circuit that consists of two series capacitor $2C$ and one shunt inductor L connected by two equal transmission lines as indicated in Figure 6.1 (c). In this case, the total actual length (ϕ) between the two adjacent CPW elements comprises up the three sections, i.e., $\phi = \phi_c + \phi_t + \phi_l$, where ϕ_t is the phase of connecting CPW lines between the two elements.

In fact, these 1-D infinite-extent CPW metamaterials can be recognized as the periodic transmission lines [54] or 1-D composite materials [60, 66, 133]. From the macroscopic point of view, they can be further considered as an effective uniform transmission line medium, whose 1-D material parameters are dispersive and complex per-unit-length transmission parameters [122, 125]. To determine these parameters, the ABCD matrix of the whole unit cell with the length, T , is at first derived via simple cascaded network. Following the discussion in Sec. 6.2, the effective propagation constant (γ) can thus be simply expressed in the closed form as the function of normalized reactance and susceptance:

$$\cos(\gamma T) = \cos 2\phi + \sin 2\phi / 2\bar{B} + \sin 2\phi / 2\bar{X} - \cos^2 \phi / 2\bar{B}\bar{X} \quad (7.1)$$

Also, the effective characteristic impedance or Bloch-wave impedance of such 1-D metamaterial can be obtained via the calculated propagation constant γ .

$$Z_c = -j \sinh(\gamma T) Z_{0u} / (\sin 2\phi - \cos^2 \phi / \bar{X}) \quad (7.2)$$

Z_{0u} is the characteristic impedance of uniform CPW as well as $\bar{X} = X / Z_{0u}$ and

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$$\bar{B} = B / Y_{0u}$$

Figure 7.6 Frequency dispersive curves of phase constant βT under combination of lumped capacitive and inductive coupling elements

As discussed in Chapter 6, there exists four regions in the considered frequency range covering the LH stopband, LH passband, LH/RH bandgap and RH passband. Particularly, in the LH passband, propagation constant exhibits purely negative imaginary value while in RH passband, it exhibits the purely positive imaginary value. The phase constant βT as a function of frequency under lumped element unit cell is plotted in Figure 7.6. It can be observed that in the LH stopband ($f < f_1$), βT keeps constant of $-\pi$, in the LH passband ($f_1 < f < f_2$), βT increases from $-\pi$ to 0, while in the middle bandgap region ($f_2 < f < f_3$), it keeps zero and then increases gradually to π in the RH passband ($f_3 < f < f_4$). These angular frequencies (ω_1 to ω_4) can be derived from equation (7.1).

The lower frequency edge ω_1 of LH passband locates where $\cos rT$ firstly reaches -1, thus,

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ω_1 will satisfy the following equation:

$$4 \cos^2 \phi \cdot \bar{L} \bar{C} \omega^2 + (\bar{L} + \bar{C}) \sin 2\phi \cdot \omega - \cos^2 \phi = 0 \quad (7.3)$$

$\bar{X} = \omega \bar{L}, \bar{B} = \omega \bar{C}$, ϕ is the total electrical length connecting two elements. So the lower edge of LH passband ω_1 is:

$$\omega_1 = \frac{-(\bar{L} + \bar{C}) \sin \phi + \sqrt{(\bar{L} + \bar{C})^2 \sin^2 \phi + 4 \bar{L} \bar{C} \cos^2 \phi}}{4 |\cos \phi| \cdot \bar{L} \bar{C}} \quad (7.4)$$

From the Figure 7.6, the upper edge of LH passband or lower edge of middle stopband ω_2 and upper edge of middle stopband or lower edge of RH passband ω_3 are located where $\cos rT=1$, thus they satisfy the equation:

$$4 \sin^2 \phi \cdot \bar{L} \bar{C} \omega^2 - (\bar{L} + \bar{C}) \sin 2\phi \cdot \omega + \cos^2 \phi = 0 \quad (7.5)$$

So,

$$\omega_2 = \frac{c \tan \phi}{2 \max(\bar{L}, \bar{C})} \quad (7.6)$$

$$\omega_3 = \frac{c \tan \phi}{2 \min(\bar{L}, \bar{C})} \quad (7.7)$$

Finally, the upper edge of RH passband ω_4 is satisfied with the equation (7.3), it locates where the total frequency related electrical length $\phi = \pi/2$, that is:

$$\omega_4 = (\pi/2 - \phi_l - \phi_c) / 2\pi \cdot \frac{c}{t \sqrt{\epsilon_{\text{eff}}}} \quad (7.8)$$

where c is the light velocity in free space, t is the transmission line length between series-C and shunt L.

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Thus, equations (7.4-7.8) can be considered as a basic guideline for different metamaterial design with turntable LH and RH passbands. Figure 7.6 depicts three sets of phase constant βT frequency responses under different lumped inductive element gap widths G_2 . By fixing the capacitive coupling dimensions, an unchangeable normalized $\bar{B}$ is determined, and also fixed the total electrical length ϕ of unit cell, the middle bandgap region (f_2 to f_3) is observed to be reduced. Furthermore, the lower edge of LH passband f_1 moves to lower frequency while the upper edge of RH passband f_4 keeps unchanged with decreased G_2 . Thus, it demonstrates an increased bandwidth for both LH and RH passbands with enlarged normalized reactance $\bar{X}$. As indicated in Figure 7.6, the middle stopband gap is still large by reducing the lumped inductive element gap to its smallest value. According to the equation (7.6) and (7.7), the bandgap can be vanished by choosing two equal normalized capacitor $\bar{C}$ and inductor $\bar{L}$. The normalized reactance $\bar{X}$ of lumped inductive element is much smaller than the normalized susceptance of lumped capacitive element, so in order to obtain wider bandwidth of LH or RH passbands, larger reactance is consequently required.

On the other hand, CPW metamaterials, consisting of lumped capacitive and distributed inductive elements are comparatively studied to further minimize the concerned bandgap. By fixing the gap G_1 as smallest value 5mil, gap width G_4 as 5mil and the total electrical length ϕ , the frequency dispersive phase constant curves and characteristic impedances under different three stub length d_2 are plotted in Figure 7.7 (a) and 7.7 (b), respectively. It can be noted that in the case of shortest stub length $d_2=20\text{mil}$, normalized reactance $\bar{X}$ is still smaller than susceptance $\bar{B}$. So, the lower edge of middle bandgap f_2 is

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determined by capacitor ($\bar{B}/\omega$) while the upper edge of bandgap f_3 is determined by normalized inductor ($\bar{X}/\omega$). In this case, the real part of characteristic

Figure 7.7 (a) Frequency dispersive curves of phase constant βT (b) characteristic impedance under combination of lumped capacitive and distributed inductive elements

impedance as noted by the thinnest solid line in Figure 7.7 (b) exhibits a quasi half ellipse

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shape in LH passband, keeps zero in bandgap. In the meantime in the RH passband, it first goes into positive infinity and then falls down to its bottom and rises up again to infinity. The imaginary part of characteristic impedance noted by the dotted line in Figure 7.7 (b) increases from negative value to zero in the LH stopband, and keeps zero in the LH passband, then it increases gradually to positive infinity during the bandgap region. In the RH passband, it keeps zero again.

As d_2 increases to 40mil, the reactance is enlarged to be equal with the susceptance and f_3 moves up to be equal with f_2 , thus a seamless connection between LH and RH passband is realized with vanished bangap region. In such a case, the real part of characteristic impedance increases from zero, at lower edge of LH passband f_1 , to its maximum value and then goes down to zero at f_2 (f_3). Afterwards, it approaches positive infinity, falls down and rises up to infinity in the frequency range of f_3 to f_4 . Meanwhile the imaginary part keeps zero in the seamless LH and RH passbands.

With further extending this stub width d_2 to 60mil, normalized inductor ($\bar{X}/\omega$) exceeds the normalized capacitor ($\bar{B}/\omega$), thus showing that f_3 is unchanged and f_2 is moved forward. The shape of characteristic impedance exhibits a great change compared to the first two cases. In the LH stopband, it is purely negative imaginary value. In the LH passband, it shows a pure real positive value and increases from zero to infinity. In bandgap region, it is pure negative imaginary value and increases from negative infinity to zero. In the RH passband, it becomes purely real value again and goes up from zero to positive infinity. Table 7.1 summarizes the phenomena discussed above under different

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eight values of distributed inductive element parameters d_2 . It indicates that the bandwidth of LH region can be increased with extended d_2 . However, we can find the enhancement becomes little after d_2 is larger than 40mil. So, if we need more bandwidth, normalized capacitor ($\bar{B} / \omega$) should be enhanced as well.

d_2 (mil)	10	20	30	40	50	60	70	80
$\bar{X}$ (6.6GHz)	0.16	0.19	0.22	0.25	0.27	0.29	0.30	0.31
$\bar{B}$ (6.6GHz)	0.25	0.25	0.25	0.25	0.25	0.25	0.25	0.25
f_1 (GHz)	5.41	5.22	5.05	4.89	4.77	4.65	4.54	4.44
f_2 (GHz)	6.60	6.60	6.60	6.60	6.47	6.32	6.19	6.08
f_3 (GHz)	7.35	7.08	6.84	6.60	6.60	6.60	6.60	6.60
f_4 (GHz)	9.32	9.32	9.32	9.32	9.32	9.32	9.32	9.32
LH BW (%)	19.8	23.4	26.6	29.8	30.2	30.4	30.8	31.2
RH BW (%)	23.6	27.3	30.7	34.2	34.2	34.2	34.2	34.2

Table 7.1 Characteristics of CPW metamaterial by combination of lumped capacitive element and distributed inductive element under 8 different lengths of d_2

7.4 Infinite Metamaterial Characterizations

In order to demonstrate physical characteristics of CPW different metamaterials with varied LH and RH bandwidths, two types of metamaterials with lumped and distributed element are constructed as shown in Figure 7.8 (a) and 7.8 (b).

Based on the discussion in previous section and Chapter 6, the complex propagation constant as well as characteristic impedance are shown in Figure 7.9 and 7.10, respectively.

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Figure 7.8 Schematics and equivalent circuit models of the two infinite-extent CPW metamaterials on the substrate of $\epsilon_r=10.8$ and $h=0.635\text{mm}$. (a) Lumped CPW elements ($T_1=8.382\text{mm}$, $G_1=G_2=0.127\text{mm}$, CPW: slot width = 0.889mm and strip width = 1.016mm ; (b) Distributed CPW elements ($T_2=7.366\text{mm}$, $G_3=G_4=0.127\text{mm}$, $l_2=0.508\text{mm}$, $d_1=0.508\text{mm}$, $d_2=2.032\text{mm}$).

Figure 7.9 (a) and 7.9(b) depict the complex propagation constant ($\gamma=\alpha+j\beta$) normalized to free space wave number (β_0) and characteristic impedance ($Z_c=\text{Re}(Z_c)+j\text{Im}(Z_c)$) of the infinite-extent CPW metamaterial with lumped elements in the frequency range, covering the LH stopband, LH passband, middle or LH/RH stopband and RH passband. Actually, these symmetrical periodic CPW structures are bidirectional, here the +sign of propagation constant γ refers to direction of forward wave +z direction, where the real part of Z_c always keep positive. In the low LH stopband, β is always negative and β/β_0 rises up as a linear function of frequency under the criteria of $\beta T = -\pi$ while α quickly falls down from the positive infinity ($+\infty$) towards zero. In this stopband, Z_c has the only

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negative imaginary quantity, i.e., $\text{Im}(Z_c) < 0$, with $\text{Re}(Z_c) = 0$ and this $\text{Im}(Z_c)$ tends to move up to zero with the frequency. As the frequency is raised into the LH passband, α suddenly becomes zero while the negative β goes up in an accelerated speed towards zero. In this band, $\text{Im}(Z_c) = 0$ while $\text{Re}(Z_c)$ increases above zero and then falls down to zero again. In the LH/RH stopband, $\alpha > 0$ and $\beta = 0$ while $\text{Re}(Z_c) = 0$ and the positive $\text{Im}(Z_c)$ continues to rise up until $+\infty$ with the frequency. As the frequency exceeds 7.2GHz, the RH passband is constituted, where $\alpha = 0$ and β increases from zero to the value $\beta T = \pi$. Meanwhile, Z_c suddenly becomes the only positive $\text{Re}(Z_c)$ again and gradually drops off with frequency.

Based on the above discussion, the negative and positive phase constants (β) in the LH and RH passbands have actually demonstrated the fundamental backward- and forward-wave characteristics with the anti-paralleled and paralleled phase and group velocities. These two distinctive passbands are sandwiched by the LH/RH stopband and this stopband sometimes needs to be eliminated for some applications [138]. Following the discussion in previous section, distributed element composed metamaterial is demonstrated to vanish such bandgap as well as increased RH and LH bandwidth. Figure 7.10 (a) and 7.10(b) illustrate the two effective per-unit-length parameters of the optimized CPW metamaterial with distributed CPW elements. In this case, we realize that the LH passband is directly connected with the RH passband at the frequency of 4.5GHz. As can be seen in Figure 7.10(a), over the entire LH/RH passband (3.2 to 6.8 GHz), β now smoothly increases across zero from the negative to the positive while $\alpha = 0$.

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Figure 7.9 (a) Complex propagation constant (γ) and (b) characteristic impedance (Z_c) of the infinite-extent CPW metamaterial with lumped elements

In this combined passband, $\text{Im}(Z_c)=0$ while $\text{Re}(Z_c)$ sharply falls down to zero as the frequency increases close to 4.5GHz (LH/RH transition) in the LH passband and then suddenly jumps to $+\infty$ beyond 4.5GHz. Afterwards, this $\text{Re}(Z_c)$ quickly drops down and then gradually goes up again.

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Figure 7.10 (a) Complex propagation constant (γ) and (b) characteristic impedance (Z_c) of the infinite-extent CPW metamaterial with distributed elements.

7.5 Finite-extent Metamaterial

In this section, the two finite-extent CPW metamaterials with 50Ω feed lines are characterized and then implemented to experimentally verify the derived per-unit-length parameters and demonstrate the actual guided-wave transmission across the truncated 1-

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D metamaterials. Figure 7.11(a) and 7.11(b) indicate the predicted and measured transmission coefficients (S_{21}) of the two 50 Ω -fed CPW metamaterials with 9 unit cells. The dotted lines are calculated results based on the truncated effective uniform CPW lines with the known Z_c and γ . The solid ones are experimentally de-embedded at the two truncated interfaces of the two fabricated finite-extent CPW metamaterials. This set of

Figure 7.11 Magnitude ($|S_{21}|$) and phase (ϕ_{21}) of transmission coefficients for the finite-extent CPW metamaterials: (a) lumped and (b) distributed elements.

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figures show us again that the two distinctive passbands are separated by the middle bandstop in the lumped case while these two passbands are directly linked together in the distributed case. Regardless, the phase (ϕ_{21}) has “+” and “-” signs in the LH and RH passbands, respectively. As illustrated in Figure 7.11(b), ϕ_{21} in the distributed case continuously and smoothly drops off across zero as a linear function of frequency over the range of 3.2GHz to 6.8GHz. However, due to the impedance mismatch at the interfaces between the finite-extent CPW metamaterial and 50 Ω CPW feed lines, multiple reflections bounce in the middle CPW metamaterial region, causing several ripples in magnitude ($|S_{21}|$) in the LH and RH passbands, as can be observed in Figure 7.11(a) and 7.11(b).

7.6 Conclusion

In this chapter, two lumped and distributed capacitive and inductive elements are firstly investigated for modelling of 1-D CPW metamaterial on a basis of equivalent circuit network. The numerically extracted normalized shunt reactance $\bar{X}$ and series susceptance $\bar{B}$ as well as their electrical length ϕ_l and ϕ_c demonstrate the varied LH/RH passbands via dispersive phase constant under different combination of these four elements in a comprehensive way. Then, infinite- and finite-extent CPW transmission line metamaterials with simultaneous loading of CPW lumped and distributed series-capacitive and shunt-inductive elements have been investigated in theory and experiment. Effective dispersive and complex propagation constant and characteristic impedance are numerically extracted via our self-calibrated MoM approach to demonstrate the

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fundamental guided-wave characteristics in the LH stopband, LH passband, LH/RH stopband and RH passband. Moreover, the two 9-cell CPW metamaterials with 50Ω feed lines are characterized as the truncated effective uniform transmission line section. Predicted transmission coefficients in magnitude and phase over the broad frequency range are well confirmed by experiment.

Chapter 8

Conclusion and Recommendations

8.1 Conclusion

Coplanar Waveguide (CPW) gains more and more research interests in modern microwave and millimeter wave integrated circuit design (MMIC) due to its attractive features over the conventional microstrip lines. These include ease in mounting series/shunt elements without needing via holes and low frequency dispersion of its dominant CPW mode.

With great development of MMIC, CPW discontinuities play more and more important role in the design of planar integrated circuit. Therefore, a good understanding of the characteristics of different types of CPW discontinuities is essential for design success. In the past few decades, great effort has been made to accurate characterization of a variety of CPW circuits and discontinuities by developing different advanced fullwave approaches, of which, the method of moments (MoM) has been recognized as the most efficient and powerful algorithm. Also, several commercial MoM simulators have been

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explored in the recent years for direct simulation and optimization of a large number of CPW circuit blocks with complex configurations. However, such a direct EM simulation requires intensive CPU time and huge memory thus preventing one from physical insight into the operation principle for the circuit design. In order to overcome the gap between numerical electromagnetic analysis and network based optimization procedure a called MoM-SOC technique is developed in this work to characterize these CPW discontinuities such that the optimization can be executed via resultant network topology.

To fulfill the goal introduced in Chapter 1, we first understood extensive literature reviews about different numerical modeling techniques for CPW discontinuities, design theorem for planar filter circuit and fundamentals of planar metamaterials with periodic configuration.

An efficient and accurate de-embedding technique (MoM-SOC) is introduced to calibrate the port discontinuity effects introduced by impressed field in the MoM formulation. The two-port CPW discontinuities are then modeled in terms of its impedance or admittance network, ie. Z- or Y- matrix.

This MoM-SOC technique is then used to numerically de-embed and extract the equivalent π - or T- type network parameters of various CPW discontinuities, including lumped/distributed capacitive/inductive coupling element and asymmetric overlap inductive coupling element and interdigital capacitive coupling element, as well as a simple uniform CPW transmission line. These basic network parameters are prepared for

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further application to design of filters and novel CPW metamaterials.

Compact quarter-wavelength CPW bandpass filters are then designed via simple cascaded network topology based on the impedance (K-)/admittance (J-) inverter that is obtained from equivalent T- or π -type circuit parameters. Moreover, transmission zeros are generated by asymmetric parallel-coupled structure around the half wavelength for harmonic-suppressed filter design. By differentiating the coupled spaces and overlapped lengths of two coupling elements, the improved two-stage and four-stage quarter wavelength bandpass filters with 2nd harmonic suppression are realized. The predicted results are validated by the fullwave simulation based on the Agilent ADS software and measurement of the fabricated circuits.

Novel ultra-wideband (UWB: 3.1-10.6GHz) bandpass filters on CPW are presented, designed and implemented by using the non-uniform or multiple-mode resonator (MMR) and tight coupling elements. First, MMR with three distinctive sections is constructed to generate the first three resonator modes occurring around lower-end, center and higher-end of UWB band. Then, CPW coupled element is characterized via J- or K-inverter to demonstrate the emergence of the two additional resonant poles below and above the central frequency. As a result, five-pole UWB bandpass filter with only one guided-wavelength is formed and its overall performances are verified by experiment measurements.

1-D CPW transmission line metamaterial with simultaneous series-capacitive and shunt-

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inductive loading in periodic interval is investigated to demonstrate the forward- and backward- wave propagation properties in the two distinctive passbands. Using the MoM-SOC approach, two effective per-unit length parameters, i.e., propagation constant and characteristic impedance are numerically extracted. As such, it allows us to explicitly derive the effective permittivity and permeability of 1-D composite metamaterials. Moreover, the experimental verification is made on the predicted two types of 9-unit cell CPW metamaterials

Infinite- and finite-extent CPW transmission line metamaterials with simultaneous loading of CPW lumped and distributed series-capacitive and shunt-inductive elements have been extensively investigated in theory and experiment to demonstrate the varied passbands performances. Effective dispersive and complex propagation constant and characteristic impedance are numerically extracted via our self-calibrated MoM-SOC approach to demonstrate the fundamental guided-wave characteristics in the LH stopband, LH passband, LH/RH stopband and RH passband. Further the LH and RH passbands are exhibited to be connected with a seamless bandgap by using CPW distributed element with tightened extent. Moreover, the two 9-cell CPW metamaterials with 50Ω feed lines are characterized as the truncated effective uniform transmission line section. Predicted transmission coefficients in magnitude and phase over the broad frequency range are well confirmed by experiment.

8.2 Recommendations for Future Research

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Although many works have been carried out to model various CPW discontinuities and design of CPW filters and periodic structures in this Ph.D thesis, there is still a large research spaces for further development in the field of modeling method and circuit design optimization. Some possible prospects are discussed as below in the future:

1. The MoM-SOC is presented in this thesis for modeling of CPW single layered discontinuities. This method can be further extended to characterize multi-layered CPW circuit and structures by newly deriving the dyadic Green's function.
2. Admittance or impedance inverter network parameters provide us with direct characterization of coupling behavior for different CPW structures. They can be utilized to the design of other types of wideband filters with specified passband and stopband performances in terms of the frequency-dispersive quantities, as discussed in Chapter 4 and Chapter 5.
3. Novel UWB filter using the multi-mode resonator (MMR) and tight coupling element is proposed to achieve a good performance in UWB passband in Chapter 5. However, lower and upper stopband performance of UWB filter needs to be improved by exploring new structures in order to meet the FCC requirements.
4. Newly proposed CPW metamaterials may be applied to design of various planar circuits and antenna, such as phase shifter, leaky-wave antenna and so on. In particular, a novel leaky-wave antenna can be developed to fulfill the end-fire to

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- broadside frequency scanning of its main radiation beam. Also the impedance matching network of finite metamaterial can be investigated for the application in ultra wideband bandpass filter with seamless connection of LH and RH passbands.
5. One of main obstacles for CPW applications is the accident excitation of unwanted coupled slotline (CSL) mode in asymmetrical CPW discontinuities such as bends and T-junctions. Until now, little research has been concentrated on the analysis of this CSL mode. As a challenging but exciting task, the mode conversion between CPW mode and CSL mode must be investigated and discussed in the future. As a result, various CPW novel circuits can be constructed with dual CPW/CSL-mode operation or good CSL-mode suppression.

Author's Publications

Journal Papers:

1. Jing Gao and Lei Zhu, "Accurate circuit models of CPW coupling elements for the design of compact quarter-wavelength band-pass filters," *International Journal of RF and Microwave Computer-Aided Engineering*, vol. 14, no. 9, pp. 453-461, Sep. 2004.
2. Jing Gao and Lei Zhu, "Investigation on asymmetric parallel-coupled CPW for $\lambda/4$ bandpass filters with broad rejection band," *IEICE Electron. Express*, vol. 1, no. 1, pp. 1-6, April 2004.
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4. Jing Gao and Lei Zhu, "Characterization of infinite- and finite-extent coplanar waveguide metamaterials with varied left- and right-handed passbands," *IEEE Microwave and Wireless Compon. Lett.*, vol. 15, no. 11, pp. 805-807, Nov. 2005.
5. Jing Gao, Lei Zhu, Wolfgang Menzel, and Frank Bögelsack, "Short-ended CPW multiple-mode resonators for UWB bandpass filters," *IEEE Microwave and Wireless Compon. Lett.*, vol. 16, no. 3, pp. 104-106, March 2006.
6. Jing Gao and Lei Zhu, "Effective complex permittivity and permeability of composite planar left-handed transmission line metamaterials," *Microwave Opt. Technol. Lett.*, accepted in March 2006.
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Conference Papers

1. Jing Gao and Lei Zhu, "Guided-wave characteristics of CPW transmission line metamaterials: effective per-unit length parameters," in *IEEE International Workshop on Antenna Technology and Novel Metamaterials*, March 2005, pp. 175-178.
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