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FULL-WAVE CHARACTERIZATION OF MICROSTRIP LINE STRUCTURES
FOR INNOVATIVE DESIGN OF MICROWAVE FILTERS

SUN SHENG

2006

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SCHOOL OF ELECTRICAL & ELECTRONIC ENGINEERING
NANYANG TECHNOLOGICAL UNIVERSITY

2006

**Full-Wave Characterization of Microstrip Line
Structures for Innovative Design of
Microwave Filters**

Sun Sheng

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in fulfilment 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 and has not been submitted for a higher degree to any other University or Institution.

Date

Sun Sheng

DEDICATION

To

My supervisor

And

My grandparents, my parents and my wife.

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First of all, I would like to express my sincere thanks to the School of Electrical and Electronic Engineering of the Nanyang Technological University (NTU) for the financial support during the whole duration of my studies here.

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Most of all, I'd like to thank my wife *Han*, for everything she's done for me; for sticking with me and putting up with my best and my worst during these years.

SUMMARY

The objective of this PhD dissertation is to originally demonstrate the guided-wave characteristics of finite-ground microstrip line (FGMSL) for application in design of size-miniaturized low-pass filters and to characterize the periodically nonuniform coupled microstrip lines (PNCML) for exploration of parallel coupled microstrip line (PCML) with harmonic suppression. To do it, a full-wave self-calibrated method of moments (MoM) is developed for accurate modeling of these transmission line structures with varied configurations.

After literature survey is summarized, the FGMSL open-end discontinuities are firstly characterized as a unified equivalent circuit model with a fringing capacitance and radiation conductance. It is done by integrating the short-open calibration (SOC) procedure into a determinant MoM, i.e. MoM-SOC technique. Extensive results are provided to exhibit that the capacitance increases as a decelerated function of the finite-ground width and length while the conductance is negligibly small as compared with its imaginary part. High-impedance property of an offset FGMSL is then modeled and utilized to make up an improved transmission line electromagnetic bandgap (EBG) structure with enhanced bandgap width and attenuation. The two effective per-unit-length parameters are extracted via MoM-SOC to reportedly show the fundamental frequency-dispersive characteristics of the guided-wave propagating in this EBG structure. In parallel, the transmission coefficients of the two finite-length EBG circuits with ideal impedance matching at two ports are characterized, thus exhibiting the distinctive bandstop and bandpass behaviors as the guided-wave propagates across this EBG with finite-extended region. A five-cell EBG circuit, fed with the standard 50Ω feed lines, is further

modeled to give an evident verification on its enhanced EBG performance. After the uniform FGMSL and its constituted circuit elements are thoroughly characterized, an innovative design of a stopband-enhanced and size-miniaturized lowpass filter (LPF) is presented. The finite-extended FGMSL section with equally widened strip/ground or offset narrow strip/ground conductors are studied in terms of equivalent T- or π -network, thereby constructing the modified microstrip line shunt capacitive or series inductive element.

Secondly, even- and odd-mode guided-wave characteristics of periodically nonuniform coupled microstrip lines (PNCML) are thoroughly investigated in terms of the two sets of per-unit-length transmission parameters i.e. characteristic impedance and phase constant. In this way, by executing the MoM-SOC technique, the two-port ABCD matrices of the PNCML with finite unit cells are numerically de-embedded via two sets of SOC standards so as to explicitly derive the effective per-unit-length parameters. After our investigation on the behaviors of numerical convergence, extensive results are derived to demonstrate the frequency- and periodicity-dependent per-unit-length parameters of the three types of PNCMLs against those of the uniform CML. As a result, frequency-dependent coupling strength between the lines of the finite-extended PNCML is exposed via two dissimilar impedances. Meanwhile, two phase constants aim at being equalized at a certain frequency by properly adjusting the slit depth and periodicity, thus creating a transmission zero in the first harmonic resonant frequency. Further, equivalent J-inverter network parameters of this finite-length PNCML are derived to reveal the relationship between the transmission zero and the concerned harmonic resonance. By allocating this zero to the frequency twice the fundamental passband, the

PNCML filters are then designed, fabricated and measured to showcase the advantageous capacity of the proposed technique in harmonic suppression.

Next, dual-band bandpass filters (BPFs) with controllable fractional bandwidths (FBWs) are proposed and constructed by cascading the multiple $\lambda/2$ stepped-impedance resonators (SIRs) through the distributed parallel-coupled microstrip line (PCML). This PCML with different overlapped lengths are studied to show their distributed coupling performance in terms of explicit J-inverter susceptances. It implies that the coupling degrees around two resonances can be adjusted with freedom to control the fractional bandwidths of these dual passbands. Finally, a compact dual-band microstrip bandpass filter is designed to operate at 2.4 and 5.2 GHz without needing any external impedance-matching block. The modified half-wavelength stepped-impedance resonator with sinuous configuration is constructed to simultaneously excite the dual resonances at these two specified frequencies with miniaturized overall size. Further, the novel microstrip dual-band bandpass filters with controllable fractional bandwidths and good in-between isolation are presented and implemented. Two types of coupled microstrip lines in the anti- and pro-parallel formats are then investigated in terms of an equivalent J-inverter network. Extensive results are derived to show their frequency-distributed coupling performances under different coupling lengths. The coupling degrees around the two resonances of these two coupled lines are tuned to realize the dual-passbands with different bandwidths. In addition, the anti-parallel coupled line is modeled to generate a transmission zero between the two resonances so as to achieve the good in-between isolation.

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LIST OF SYMBOLS

E^i	Impressed electric field
E^s	Scattered electric field
f	Normal frequency
$\overline{\overline{G}}$	Spatial dyadic Green's functions
$\hat{\overline{\overline{G}}}$	Spectral dyadic Green's functions
h	Thickness
H^i	Impressed magnetic field
H^s	Scattered magnetic field
J^{inc}	Impressed electric current
J_s	Surface electric currents
$\vec{J}$	Current density
k_0	Free-space wave number
$\vec{k}_t$	Transverse vector wave number
M^{inc}	Impressed magnetic current
M_s	Surface magnetic currents
$\hat{n}_i$	Longitudinal direction
r^+, r^-	Field points at the inner and outer surface of the object
W, II, I_V, V_I	Modal amplitudes, voltages and currents regarding to the dyadic spectral eigenfunctions
X_i	Error terms in error box
Z_0	Characteristic impedances

$\hat{z}$	Outward unit normal to the strip conductor surface
α	Attenuation constant
β	Phase constant
ϵ_0	Permittivity of vacuum
ϵ_{eff}	Effective dielectric constant
ϵ_r	Relative permittivity
γ	Propagation constant
λ	Wavelength
μ_0	Permeability of vacuum
μ_r	Relative permeability
$\rho = \hat{x}x + \hat{y}y$	Projection of r on the (x, y) plane
ω	Angular frequency
$\vec{\Gamma}_k, \bar{\Gamma}_k$	Reflection coefficients at the reference plane looking into right- and left-side
$\vec{Y}_k, \bar{Y}_k$	Input admittances at the reference plane looking into right- and left-side

LIST OF ABBREVIATIONS

3-D	Three-dimensional
BPF	Bandpass filter
CBCPW	Conductor-backed coplanar waveguide
CML	Coupled microstrip line
CPS	Coplanar stripline
CPW	Coplanar waveguide
CSRR	Complementary split-ring resonator
DGF's	Dyadic Green's functions
EBG	Electromagnetic bandgap
EFIE	Electric-field integral equation
EM	Electromagnetic
EME	Electric-magnetic-electric
E.W.	Electric wall
FBW	Fractional bandwidth
FDTD	Finite difference time domain
FGMSL	Finite-ground microstrip line
HFSS	High-frequency structure simulator
LNA	Low-noise amplifier
LPF	Lowpass filter
LTCC	Low-temperature cofired ceramics
MIC	Microwave integrated circuit
MMIC	Monolithic microwave integrated circuit

MoM	Method of moments
MSL	Microstrip line
M. W.	Magnetic wall
PBG	Photonic bandgap
PCML	Parallel coupled microstrip line
PNCML	Periodically nonuniform coupled microstrip line
PWS	Piecewise sinusoidal
RF	Radio frequency
SIR	Stepped-impedance resonators
SISO	Single-input-single-output
SOC	Short-open calibration
SRR	Split-ring resonator
TE	Transverse-electro
TEM	Transverse-electro-magnetic
TM	Transverse-magnetic
TL	Transmission line
TLGF	Transmission-line Green's function

CHAPTER 1

Introduction

1.1 BACKGROUND

The frequency range of *microwave* is from 300 MHz to 300 GHz, with a corresponding electrical wavelength from 1m to 1mm in free space. Over the past one hundred years, microwave has become a ubiquitous technology that is used not only for radars and communication systems, but also for the characterization and analysis of materials, cooking and industrial drying, medical diagnosis and treatment, radio astronomy, and so on [1]–[2].

Since the 1950s, the emergence of microwave integrated circuits (MICs), especially monolithic microwave integrated circuits (MMICs), has been playing a very important role in the advanced radio frequency (RF) & Microwave technologies [3]–[5]. Many advantages have been recognized and well known in MICs/MMICs, such as compact size, light weight, high integration, low cost and so on. As the fundamental structures in MICs, planar transmission lines, such as

Fig. 1.1. Cross-sectional views of planar transmission lines for MICs. (a) Stripline. (b) Microstrip line (MSL). (c) Coupled microstrip line (CML). (d) Coplanar stripline (CPS). (e) Finline. (f) Slotline. (g) Conductor-backed coplanar waveguide (CBCPW). (h) Coplanar waveguide (CPW).

stripline, *microstrip line* (MSL), *coupled microstrip line* (CML), slotline, coplanar waveguide (CPW), coplanar stripline (CPS), etc., are used for realizing the basic blocks for MICs [5]–[8]. Fig. 1.1 shows the cross-sectional view of these lines. These planar structures are the backbone of MICs, and offer an efficient and flexible way in the design of MICs.

The MSL concept was first proposed in 1952 by Greig, Englemann, Assadourian, Rimai, and Kostriza of ITT Federal Telecommunications Laboratories [1], [9]–[11]. It employs a flat strip conductor suspended above a ground plane by a low loss dielectric material, as shown in Fig. 1.1(b). As usual, the width of the ground plane is assumed electrically wide or infinite. Thus, the microstrip circuits with different electrical properties can be only realized relying on the geometrical configuration of the strip conductor. Initially, the center conductor of coaxial line was flattened into a strip and the outer conductor was altered into a rectangular box, making itself more

suitable for the fabrication of components. Subsequently, stripline, as shown in Fig. 1.1(a), came out by removing the side walls. This type of transmission line was developed very quickly and is still in use today due to its two good technical properties: (1) all discontinuity elements of the center strip are purely reactive; (2) the characteristic impedance and the phase velocity (or effective dielectric constant) of the dominant (TEM: transverse-electro-magnetic) mode are independent of the frequencies [1].

Unlike stripline, MSL appeared by removing both side walls and top plate, leaving only the strip and the bottom plate with a dielectric substrate layer between them to support the strip. Fig. 1.2 shows the three-dimensional (3-D) geometry of MSL. Most of its field lines or energies concentrate in the dielectric region underneath the strip conductor except that a small fraction of the fields are scattered in the air

Fig. 1.2. Three-dimensional (3-D) geometry of microstrip line.

region. For these reasons, it cannot support the purely TEM wave, but the dominant mode still can be considered as the quasi-TEM mode at low frequencies. All discontinuity elements became resistive and have radiation. Meanwhile, the characteristic impedance and the phase velocity became the frequency-dependent parameters. Since then, MSL has become the best known and most widely used planar transmission line for RF and microwave integrated circuits, due to its planar nature, ease of fabrication using photolithographic processes, easy integration with

solid state devices, good performance of heat sinking and mechanical support, as well as vast design information [12]–[14].

1.2 MOTIVATION

1.2.1 FINITE-GROUND MICROSTRIP LINE

An alternative microstrip line with finite-width substrate was first presented by Smith and Chang [15], in 1980. This structure seems more close to the practical case and its geometry is described in Fig 1.3(a) as a transmission line with a finite-width dielectric substrate and infinite-width ground as compared with traditional counterpart in Fig. 1.2. In 1985, the modified structure of microstrip structure with both finite-width dielectric and finite-width ground plane was considered by Smith and Chang [16], in which the strip conductor and finite ground were treated as a printed twin-line of unequal widths in an inhomogeneous media, as shown in Fig.1.3 (b). The numerical results in [16] have shown us that high characteristic impedance of the line can be achieved with the use of finite-width dielectric and ground plane. A design example (using case in Fig. 1.3(b)) was given to design a balun circuit [16]. In the past, the edge effect of the dielectric and ground plane was taken into account, in which the ground plane was treated as an additional strip conductor [17]. Meanwhile, using the method of moments, the far-field radiation for the case in Fig.

Fig. 1.3. 3-D geometry of different MSLs. (a) MSL with only finite-width dielectric substrate. (b) MSL with finite-width dielectric and ground plane.

1.3(b) was investigated by See and Freeman [18].

On the other hand, Gurel and Chew [19] studied the MSL with the finite-width ground plane printed on an infinite-width substrate in 1988. The 3-D geometry of this finite-ground microstrip line (FGMSL) is shown in Fig. 1.4. They analyzed the

Fig. 1.4. The 3-D geometry of MSL with only finite-width ground plane.

effective dielectric constant ϵ_{eff} of MSL with equal widths of strip conductor and ground plane using a 2-D method. Recently, some experimental results for FGMSL circuits on Si Wafers were reported by Ponchak, Margomenos and Katehi [20]. The measured results showed that as the ground width is reduced to three times of its strip conductor width, this FGMSL could achieve low loss on 2 Ω -cm Si. Coupling between the FGMSL embedded in 3-layer was considered in [21]. The coupling is found higher than that for infinite width ground plane, but it can be significantly reduced by connecting the two ground planes with metallic via posts.

In the past several decades, the freedom of finite-ground microstrip structure has been introduced to the design of antennas and antenna arrays. Due to the finite dimension, the diffraction from the ground plane edges will have an additional influence on the antenna properties [22]–[27]. The back radiation effect is described by [28] and the antenna's performance is optimized by adjusting the size and shape of the ground plane [29]–[31].

Very recently, the promising multilayered fabrication technologies, such as Low-temperature Cofired Ceramics (LTCC), have been directing an increasing interest in the design of size-miniatured and high-quality hybrid integrated circuits and systems for wireless communication. As a result, it needs to be promptly addressed to analyze and optimize many of FGMSL structures using the advanced analysis and design techniques. Research on multilayer structure becomes more and more important. In 1989, the multilayer dielectric structures with a ground plane were considered by Harokopus and Katehi based on the full-wave Method of Moments (MoM) [32], [33]. And many of other works have also been done for such structures as listed in [34]–[41]. But all these works cannot be directly used for the structure with finite-width ground plane. To address this issue, we focus on the modeling of multi-conductor structures on the multilayered dielectric substrates, in which the ground plane is also considered as the finite-width conductor. This is a more general case containing large design flexibility, which can be applied for the LTCC circuits more widely [42], [43], [44], and make the design more accurate and flexible. Fig. 1.5 shows cross sectional views of the multi-conductor structures embedded in multilayered dielectric substrates with infinite-width and finite-width ground plane, respectively.

Based on the above discussion, FGMSL is an improved microstrip transmission

Fig. 1.5. Cross sectional views of multi-conductor embedded in multilayered dielectric substrates. (a) With infinite-width ground plane. (b) With finite-width ground plane.

line with an additional degree of freedom using its finite ground. Such a novel structure has great potential in the hybrid MIC's with multilayered configuration. Numerical analysis of finite-width MSL has been reported to a certain degree to show its good performance including high characteristic impedance, miniaturized-size, better flexibility, etc. However, only few studies have been concentrated on the characterization of many circuit structures formulated on this FGMSL. Full-wave analysis of finite ground microwave integrated circuits/discontinuities is definitely an urgent work at present in order to apply this finite-ground MSL in the practical design.

1.2.2 PERIODICALLY NONUNIFORM COUPLED MICROSTRIP LINE (PNCML)

Coupled microstrip line (CML), as shown in Fig. 1.1(c), has been gaining a wide application in the bandpass filter design due to its distributed and tightened coupling behaviors [6], [45], [103], [104]. The 3-D geometry of this CML is shown in Fig. 1.6. By enforcing the even and odd excitations on two upper strips of the CML, as

Fig. 1.6. The 3-D geometry of CML.

shown in Fig. 1.7, two dominant modes come into being: even- and odd-mode. Due to the inhomogeneous dielectric medium, these two dominant modes exist in this CML with different velocities of propagation. This nonsynchronous feature deteriorates the performances of microstrip circuits using the uniform CML, such as

Fig. 1.7. Even and Odd excitations for the CML. (a) Even mode. (b) Odd mode.

low directivity in directional coupler [46] and spurious harmonic passband in bandpass filter (BPF) [47]. To address this issue, two effective approaches have been explored by equalizing the phase velocities and differentiating the traveling routes of even- and odd-mode.

In [48], [122], [123], over-coupled structures were constituted at the stages to extend the odd-mode phase length, thus compensating the difference in phase velocity between two modes. This way, spurious-responses at $2f_0$ [48], [122], and $3f_0$ [123] can be suppressed by adjusting the dimensions of the attached over-coupled sections. Due to the bandstop behavior, the uniplanar compact photonic bandgap (PBG) structure was integrated as ground plane to reject the spurious passband [132], [94]. The strip conductor of MSL was replaced by the electric-magnetic-electric (EME) surfaces in [133]. The split-ring resonators (SRRs) and complementary split-ring resonators (CSRRs) are attached at the two sides of the parallel strip conductors [118], [134].

On the other hand, the stepped-impedance resonator (SIR) technique was proposed in [128] and it provides us with an alternative degree of freedom in controlling the frequency resonance responses, thereby effectively widening the stopband between the dominant and 1st spurious passbands in the filter design. By

directly tap-feeding the half-wavelength transmission line resonators, the 1st spurious passband of the designed filter can also be suppressed with the emergence of the transmission zeros, as originally implemented in [135], [136].

In recent years, much more interest has been aroused to utilize the periodically nonuniform coupled microstrip line (PNCML) with various unit-cell configurations as a simple and effective approach to address this problematic issue. The strip-width modulation technique is developed to make up the “wiggle-line” BPF with good harmonic suppression over a wide frequency range [49], and the multi-spurious rejection up to $5f_0$ [131]. In parallel, the corrugated coupled microstrip line [47] was presented to extend the actual traveling path for the odd-mode such that the 1st harmonic can be suppressed by equalizing the two phase velocities of the dominant even and odd modes. Similarly, in order to achieve the same goal, the square grooves [50] were periodically etched on the parallel-coupled lines and the coupled meander lines were constituted [51]. In addition, the concerned harmonic passband can be effectively suppressed by suspending the dielectric layer above the ground plane or forming a backside aperture as discussed in [52]–[54], [121].

However, without knowing the even- and odd-mode characteristic impedances and phase velocities of these PNCMLs in advance, all the above filters can only be designed in the cut-and-try way by numerically simulating their overall layouts with varied periodicities and unit-cell dimensions toward maximizing the return loss around the harmonic passband. In order to investigate these PNCMLs in depth for circuit design, it is commonly recognized as the most critical issue to characterize the fundamental per-unit-length transmission parameters of even- and odd-mode, i.e. *phase constants and characteristic impedances*.

In [111], the spectral-domain approach is applied to analyze the effective dielectric constants of the even- and odd-mode in the PNCML. But, it does not take into account the frequency dispersion so that this approximate approach may not be suitable for efficient network-based analysis and design of microwave circuits. Very recently, the finite difference time domain (FDTD) technique is employed to analyze such two frequency-dependent phase constants of PNCML [112]. But, unfortunately, no reported work to date has been carried out to calculate the even- and odd-mode effective characteristic impedances of the PNCML with any shaped configuration.

1.2.3 DUAL-BAND MICROSTRIP BANDPASS FILTER

Multi-band devices, as shown in Fig. 1.8, such as multi-band antenna [137]–[142], multi-band filter [57]–[62], multi-band coupler [145], [146] and multi-band low-noise amplifier (LNA) [55], [143], have been recently receiving a tremendously increasing attention in exploring many advanced wireless systems with simultaneous operations at multiple frequency bands [55], [144]. As pointed out in [56], multi-band passive circuits basically determine the multi-band operation quality, overall size and fabrication cost of a RF and wireless module in the integration technologies. Of them, bandpass filter with multiple passbands is considered as one of the most key components in these multi-band systems. Due to the shortage of matured design procedure, it becomes the most challenging issue for anyone to design the multi-band filters with good passband performances even for a dual-band case.

Fig. 1.8. Multi-band devices.

At the beginning stage, the two filtering circuits with different passbands at the center frequencies f_1 and f_2 , are usually connected together to implement the initial type of a dual-band filter [57], as shown in Fig. 1.9. As a two-input-two-output structure, the LC matching circuits are required to create dual passbands at two separate frequencies. However, this solution increases the fabrication cost, the insertion loss and the overall size of a resultant filter block.

Fig. 1.9. Initial idea of dual-band filter.

To reduce the size, the single-input-single-output (SISO) concept is proposed in [58], where the two filters share one layout only, and photonic bandgap (PBG) surface is employed to add a transmission zero above the second passband. However, the return loss is still low in both passbands and size still needs to be reduced.

Recently, the stepped-impedance resonator (SIR) is utilized to make up a filter with dual passbands [59], [60]. To strengthen the Q factors or lower the insertion losses in both passbands at f_1 and f_2 , the two dual-band impedance transformers have to be additionally constructed at the two ports of the core cascaded-resonator section as implemented in [60], [61]. Thus, these dual-band distributed transformers not only significantly enlarge the overall size of the resultant filter block and also bring out the complexity in simultaneously achieving the two operating frequencies, reducing the insertion losses and adjusting their fractional bandwidths (FBWs) in both dual passbands.

To improve the stopband attenuation, coupled resonator pairs are proposed in [62] to generate two transmission zeros in the higher side of each passband. Unfortunately, the filtering performances in the dual passbands are found not so good because of the unexpected parasitic effects and approximate modeling. Based on the above discussion, it is urgent to reduce the overall size and improve the filtering performance of dual-band filter blocks.

1.3 OBJECTIVES

The goal of this dissertation is to develop a full-wave modeling for the characterization of a variety of microstrip lines and coupled microstrip lines circuits for the design of modern microwave integrated circuits, thus providing the design guideline and library for its application in the exploration of next-generation of low-loss, small-size, and good-quality hybrid RF, microwave and millimeter-wave integrated circuits. The objectives of our research can be divided into the following three parts:

1.3.1 FINITE-GROUND MICROSTRIP LINE

- To develop a full-wave model for characterization of *finite-ground microstrip line*.
- To characterize *finite-ground microstrip line* open-end discontinuity for quantitative investigation of the diverse finite-ground effects in the FGMSL structure.
- To analyze the high-impedance property of offset *finite-ground microstrip line* for enhancement of electromagnetic bandgap performance.
- To apply the high-impedance property of offset *finite-ground microstrip line* as an improved series inductive element for harmonic-suppressed lowpass filter design.

1.3.2 PERIODICALLY NONUNIFORM COUPLED MICROSTRIP LINE

- To build up a full-wave model for explicit representation of *periodically nonuniform coupled microstrip lines*.
- To characterize the fundamental per-unit-length transmission parameters of *periodically nonuniform coupled microstrip line* for two dominant modes.
- To predict the location of transmission zero for the finitely-extended *periodically nonuniform coupled microstrip line* section and further allocate it suitably to suppress the 1st-harmonic passband in the design of microstrip bandpass filters.
- To accurately reallocate the transmission zero for suppression of the 1st-harmonic passband in filter design by using frequency-dispersive J-inverter network.

1.3.3 PARALLEL COUPLED MICROSTRIP LINE

- To model the coupling dispersion of the *parallel-coupled microstrip lines* and achieve the adjustable coupling degrees at two specified frequencies for design of dual-band filters with controllable fractional pass bandwidths.
- To build up a miniaturized dual-band filter without needing external dual-band impedance transformers.
- To investigate the *parallel and anti-parallel coupled microstrip lines* for the design of advanced dual-band filter.
- To make up a novel microstrip dual-band filter with miniaturized size, controllable dual-passband widths and good in-between isolation.

1.4 MAJOR CONTRIBUTIONS

The research outcome achieved in this PhD dissertation represents the innovative applications of uniform and non-uniform microstrip lines and coupled microstrip lines in the development of modern microwave integrated circuits. The major contributions of this dissertation are listed as follows:

- ✓ Characterized the finite-ground microstrip line (FGMSL) open-end discontinuities via a self-calibrated MoM as a unified circuit model with a fringing capacitance and radiation conductance. By integrating the short-open calibration (SOC) procedure into a determinant MoM, the model parameters are extracted without needing the alternative port impedance. Regardless of non-ideal voltage sources, extracted parameters are observed to achieve a stable convergence as the feeding line is sufficiently extended.

- ✓ Utilized the high-impedance property of an offset finite-ground microstrip line to make up an improved transmission line electromagnetic bandgap (EBG) structure with enhanced bandgap width and attenuation. Using the self-calibrated method of moments, the two effective per-unit-length parameters are originally extracted to demonstrate the fundamental frequency-dispersive characteristics of the guided-wave in this EBG.
- ✓ Characterized the uniform FGMSL and its constituted circuit elements via self-calibrated MoM for innovative design of a stopband-enhanced and size-miniaturized lowpass filter (LPF). An offset FGMSL is modeled and extracted to quantitatively exhibit its capacity in achieving high characteristic impedance compared to the infinite-ground microstrip line counterpart. The finite-extended FGMSL section with equally widened strip/ground or offset narrow strip/ground conductors is then studied in terms of equivalent T- or π -network, thereby constructing the modified microstrip line shunt capacitive or series inductive element. By making effective use of enlarged series inductance of an offset FGMSL with shorter electrical length, two novel LPF blocks are proposed, designed and fabricated.
- ✓ Characterized the periodically nonuniform coupled microstrip line (PNCML) loaded with transverse slits as an equivalent uniform coupled transmission line via self-calibrated MoM technique. Full-wave modeling has been executed to investigate even- and odd-mode guided-wave characteristics of PNCML in terms of the two sets of per-unit-length transmission parameters i.e. characteristic impedance and phase constant. By applying the SOC procedure in the MoM platform, the two-port ABCD matrices of the

PNCML with finite unit cells are numerically de-embedded via two sets of SOC standards so as to explicitly derive the effective per-unit-length parameters.

- ✓ Exposed the frequency-dependent coupling between the lines of the finite-extended PNCML and equalized the two phase constants at a certain frequency by properly adjusting the slit depth and periodicity, aiming at realizing the transmission zero. By allocating this zero to the frequency twice the fundamental passband, one-stage and two-stage PNCML bandpass filters (BPFs) are then designed and fabricated with the suppressed 1st-harmonic.
- ✓ Studied the two-port PNCML with finite length an equivalent J-inverter network for microstrip bandpass filters with harmonic suppression. Extracted J-inverter susceptance is exhibited to vary up and down versus frequency with the null or transmission zero when the coupled length becomes half-wavelength. This transmission zero can be adjusted by varying the periodicity and/or slit depth to suppress the 1st spurious harmonic passband of the filter. Two one-stage and a three-stage BPFs are designed and fabricated on a basis of this technique.
- ✓ Constructed the dual-band bandpass filters with controllable fractional bandwidths by cascading the multiple $\lambda/2$ stepped-impedance resonators through the distributed parallel-coupled microstrip lines. By suitably choosing the aspect ratio of two strip widths or impedances in the SIR, the first two resonant frequencies are allocated to 2.4 and 5.2GHz for dual-band filter application. The parallel-coupled microstrip lines with different overlapped lengths are investigated to show their distributed coupling

performance in terms of explicit J-inverter susceptances. It implies that the coupling degrees around two resonances can be adjusted with freedom to control the FBW of these dual pass bandwidths. Three one-stage and three two-stage dual-band BPF circuits are designed and fabricated.

- ✓ Proposed and designed a compact dual-band microstrip BPF to operate at 2.4 and 5.2 GHz without needing any external impedance-matching block. The modified half-wavelength stepped-impedance resonator with sinuous configuration is constructed to simultaneously excite the dual resonances at these two specified frequencies with miniaturized overall size. The fabricated dual-band filter has not only significantly miniaturized the overall size, but also achieved the good dual-passband filtering performances at the two specified bands, i.e. 2.4 and 5.2 GHz.
- ✓ Presented and implemented the novel microstrip dual-band BPFs with controllable FBWs and good in-between isolation. A half-wavelength stepped-impedance resonator is firstly characterized, aiming at producing the two resonant frequencies at 2.4 and 5.2 GHz. Two types of coupled microstrip lines in the parallel and anti-parallel formats are then investigated in terms of unified equivalent J-inverter network. The coupling degrees of these two coupled lines at the two resonances are properly adjusted to achieve the dual-passband response with varied or tunable bandwidths. The parallel coupled line is modeled to bring out a transmission zero between the two resonances so as to achieve the good in-between isolation. Three two-stage bandpass filters are initially designed and one three-stage dual-band filter is in final optimally designed and fabricated.

1.5 ORGANIZATION

This PhD dissertation is organized as follows. In Chapter 2, the formulation of the integral equations for perfect strip conductors embedded in a layered medium is outlined. The dyadic Green's functions (DGF's) in the spectral domain are formulated by using a simple cascaded transmission line network in conjunction with multi-layered environment. The full-wave MoM in the spectral domain and the SOC technique self-contained in the MoM algorithm are briefly introduced.

Chapter 3 introduces the configurations of various FGMSL discontinuities and discusses the different excitation principles in order to formulate the source-types or determinant MoM algorithm. Then, some initial results about the FGMSL open-end circuit and periodic EBG FGMSL are given.

Chapter 4 investigates the offset FGMSL with narrow strip/ground width to show its enhanced characteristic impedance. By forming the shortened offset-FGMSL with high impedance, a new class of LPFs with miniaturized-size and enhanced-stopband are then proposed, designed and realized. Some design examples of LPFs are given and confirmed by the HFSS-based full-wave simulation and measurement.

Chapter 5 thoroughly characterizes PNCML as an equivalent uniform coupled transmission line via self-calibrated MoM-SOC technique. Both two effective per-unit-length transmission parameters, i.e. phase constants and characteristic impedances, are extracted to demonstrate the frequency-dispersive and periodicity-dependent guided-wave characteristics of the two dominant propagating modes.

Chapter 6 presents a systemic approach to predict the location of transmission zero of the finitely-extended PNCML section. After that, this zero is suitably

allocated to suppress the 1st-harmonic passband in the design of microstrip bandpass filters. Next, the PNCML with finite length is investigated in terms of a generalized coupling-oriented J-inverter network. It does exhibit the basic principle of suppressing the 1st-harmonic passband in filter design with the use of the transmission zero reallocation technique.

Chapter 7 discusses the coupling dispersion of the PCML with varied lengths. Some SIR dual-band filters are designed with varied dual-band FBWs. A compact design is finally implemented without needing any external dual-band impedance-transformer feeds.

Chapter 8 presents a novel dual-band bandpass filter with controllable fractional bandwidths and good in-between isolation in the middle stopband. The frequency-dependent properties of the parallel and anti-parallel coupled microstrip lines are thoroughly investigated to quantitatively demonstrate their distributed coupling performance in terms of extracted J-inverter network parameters. Furthermore, the efficient optimization design of the whole filter structures is carried out based on the simple cascaded network topology.

Finally, conclusions and recommendations are made in Chapter 9.

CHAPTER 2

Techniques in Theoretical Modeling

2.1 INTRODUCTION

In order to analyze and design various planar integrated circuits and antennas, it has been considered as a starting point to study the multilayered structures with different static and dynamic electromagnetic techniques. Rather than other numerical techniques, the method using the dyadic Green's functions (DGF's) has been found the most powerful, in which the field can be found efficiently in terms of a set of given sources. Extensive research has been carried out to formulate the generalized DGF's for layered media [63]-[74] using different analytical approach. A simple but effective way to formulate DGF's in the spectral domain, so-called transmission-line Green's function (TLGF), was originally presented by Itoh [75].

For modeling the unbounded planar structures without ground plane, the spectral-domain TLGF in the multilayered media need to be at first formulated in the open

environment. This TLGF is then used in the formulation of integral equations, which can be solved by the method of moments (MoM) [76], [77], [147]–[151]. So far, the MoM formulated in the spectral domain or space domain has been formulated and applied for comprehensive treatment of various complex planar structures, i.e. bounded/unbounded structures of numerous planar microwave passive circuits and antennas [78], [33], [79]–[85], [86], [148], [149], [151].

On the other hand, a novel numerical de-embedding technique, namely, short-open calibration (SOC) technique, was introduced by Zhu in [87] towards removing all the unwanted parasitics brought by the approximation of the impressed source in the formulation of a determinant MoM algorithm. This scheme can be directly accommodated in a full-wave MoM algorithm to extract the equivalent circuit model parameters of circuits/discontinuity elements involved in the complicated circuit layout, thus providing an efficient optimization design of the whole circuit structure in terms of network-connected topology. So far, various unbounded integrated discontinuities have been successfully analyzed by integrating this SOC with the MoM technique as done in [87], [88], [89].

2.2 SPECTRAL DYADIC GREEN'S FUNCTIONS

In 1980, Itoh [75] developed the spectral Green's functions using a simple cascaded transmission line system for modeling of planar transmission line structure, named as the immittance approach. Since then, numerous researchers have studied the spectral Green's functions for layered structures. In 1994, Pan and Wolff [72] presented a systematic approach for deducing the spectral DGF's with all the nine elements. In this method, spectral fields can be expressed by the equivalent

voltages and currents on the transmission-line analog of the medium, thus it is called as TLGF.

2.2.1 INTEGRAL EQUATIONS FORMULATION

We start to consider an arbitrarily shaped object embedded in a layered medium and excited by the known electric and magnetic currents (J^{inc}, M^{inc}) , as described in [74]. The total electric and magnetic field produced by the surface currents (J_s, M_s) and (J^{inc}, M^{inc}) can be in general expressed as:

$$(E, H) = (E^i + E^s, H^i + H^s) \quad (2.1)$$

where (E^i, H^i) are the impressed fields due to (J^{inc}, M^{inc}) , and (E^s, H^s) are the scattered fields due to (J_s, M_s) . In the linear media, the scattered fields due to (J_s, M_s) can be at first derived with the help of spatial DGF's:

$$E^s(r) = \int_V \overline{\overline{G_{EJ}}}(r, r') \cdot J_s(r') dr' + \int_V \overline{\overline{G_{EM}}}(r, r') \cdot M_s(r') dr' \quad (2.2a)$$

$$H^s(r) = \int_V \overline{\overline{G_{HJ}}}(r, r') \cdot J_s(r') dr' + \int_V \overline{\overline{G_{HM}}}(r, r') \cdot M_s(r') dr' \quad (2.2b)$$

where $\overline{\overline{G_{PQ}}}(r, r')$ is the spatial DGF, in which the subscript P indicates the induced E- or H- field at the field point r , and Q is the electric or magnetic current density at source point r' . The DGF must satisfy the dyadic version of Maxwell's equations [72].

On the surface of the two connected media in an object, the tangential components of the electric field and the magnetic field must be continuous. For the concerned microstrip line structure in this work, they must satisfy the boundary conditions over the strip conductor surface as indicated as below.

$$\hat{Z} \times (E^i(r) + E^s(r)) = 0 \quad (2.3a)$$

$$\hat{Z} \times (H^i(r^+) + H^s(r^+)) = \hat{Z} \times (H^i(r^-) + H^s(r^-)) \quad (2.3b)$$

where r^+ and r^- are the field points at the inner and outer surface of the object, $\hat{Z}$ is the outward unit normal to the strip conductor surface.

2.2.2 SPECTRAL DYADIC GREEN'S FUNCTION

The following two equations indicate the Fourier transformation and its reverse counterpart, respectively.

$$F\{A(\rho, z, \rho', z')\} = \tilde{A}(\rho, z, \rho', z') = \int \int_{-\infty}^{+\infty} A(\rho, z, \rho', z') e^{j\bar{k}_t \cdot (\rho - \rho')} dx dy \quad (2.4a)$$

$$F\{\tilde{A}(\rho, z, \rho', z')\} = A(\rho, z, \rho', z') = \frac{1}{4\pi^2} \int \int_{-\infty}^{+\infty} \tilde{A}(\rho, z, \rho', z') e^{-j\bar{k}_t \cdot (\rho - \rho')} dk_x dk_y \quad (2.4b)$$

where $\rho = \hat{x}x + \hat{y}y$ is the projection of r on the (x, y) plane and $\bar{k}_t = \hat{x}k_x + \hat{y}k_y$ is the transverse vector wave number.

Applying the above Fourier transform relations in (2.4) to the dyadic Maxwell's equations, we can obtain the dyadic version of Maxwell's equations in the spectral domain [72]. After the vector and dyadic spectral eigenfunctions for one-dimensionally inhomogeneous media are defined, the scalarized spectral DGF's can be determined from two sets of coupled transmission-line equations (2.5).

$$\begin{cases} \frac{d}{dz} W(z, z') + jk_z(z) Z(z) I_v(z, z') = 0 \\ \frac{d}{dz} I_v(z, z') + jk_z(z) Y(z) W(z, z') = \delta(z - z') \end{cases} \quad (2.5a)$$

$$\begin{cases} \frac{d}{dz} II(z, z') + jk_z(z)Z(z)V_I(z, z') = 0 \\ \frac{d}{dz} V_I(z, z') + jk_z(z)Y(z)II(z, z') = \delta(z - z') \end{cases} \quad (2.5b)$$

where W , II and I_V , V_I are modal amplitudes, voltages and currents regarding to the dyadic spectral eigenfunctions, the other coefficients are denoted by,

$$k_z(z) = \sqrt{\omega^2 \mu_0 \mu(z) \varepsilon_0 \varepsilon(z) - k_t^2} \quad (2.6a)$$

$$Z^{TE}(z) = \frac{1}{Y^{TE}(z)} = \frac{\omega \mu_0 \mu(z)}{k_z(z)} \quad (2.6b)$$

$$Z^{TM}(z) = \frac{1}{Y^{TM}(z)} = \frac{k_z(z)}{\omega \varepsilon_0 \varepsilon(z)} \quad (2.6c)$$

2.2.3 MULTILAYER STRUCTURES

A general analysis for the transverse electric point source at arbitrary location in inhomogeneous multilayer structures is at first presented here. As described in [90], the multilayer media structure in Fig.2.1 is assumed to be infinite in the x-y plane with infinitesimal conductor thickness. Each layer has its own thickness h_i , and relative permittivity and permeability ε_{ri} and μ_{ri} , respectively. The region 0 means an open medium in the upper or lower half space. In this chapter, attention is focused on considering the fully unbounded open structure in relevance to the FGMSL structure, as illustrated in Fig. 2.1(a). Fig. 2.1(b) and 2.1(c) show the half open structure for MSL and the shield structure for stripline. They can be considered as the two special cases of the fully opened structure in Fig. 2.1(a) and their relevant fields/equations can be easily derived by enforcing that the terminal admittance be zero on the shielded ground planes, i.e. $Y_0=0$.

Fig. 2.1. Generic multilayer media structures. (a) Open (b) Half open (c) Shielded.

Fig. 2.2. Equivalent transmission Line Model for multilayer Media excited by a three-dimensional electric point source.

Fig. 2.2 depicts the equivalent transmission line model for this inhomogeneous multilayer medium. To find the TLGF, the unit sources $I=1A$ and $V=1V$ are assumed to be located in the layer i at position z' . The field point is located in the interface of layers. $\bar{\Gamma}_k$, $\bar{Y}_k$ are the reflection coefficients and input admittances at the reference plane looking into right-side, while $\bar{\Gamma}_k$, $\bar{Y}_k$ are those looking into the left-side.

Here, the modal amplitudes can be expressed as a response of the field point at the right side of source as

$$W^{TE}(z, z') = \begin{cases} V^{TE}(z') & z = z' \\ V^{TE}(z') \cdot V^{TE}(z_j) & z = z_j \end{cases} \quad (2.7a)$$

$$W^{TM}(z, z') = \begin{cases} V^{TM}(z') & z = z' \\ V^{TM}(z') \cdot V^{TM}(z_j) & z = z_j \end{cases} \quad (2.7b)$$

$$H^{TM}(z, z') = \begin{cases} I^{TM}(z') & z = z' \\ I^{TM}(z') \cdot I^{TM}(z_j) & z = z_j \end{cases} \quad (2.7c)$$

where

$$V^m(z') = \frac{1}{\bar{Y}_{z'}^m + \bar{Y}_{z'}^m}, \quad I^m(z') = \frac{1}{\bar{Z}_{z'}^m + \bar{Z}_{z'}^m} \quad (2.8a)$$

$$V^m(z_j) = \frac{1}{e^{-\gamma_i h_i^+} + \bar{\Gamma}_i e^{\gamma_i h_i^+}} \cdot \prod_{k=i+1}^j \frac{1 + \bar{\Gamma}_k}{e^{-\gamma_k h_k} + \bar{\Gamma}_k e^{\gamma_k h_k}} \quad (2.8b)$$

$$I^m(z_j) = \frac{1}{e^{-\gamma_i h_i^+} - \bar{\Gamma}_i e^{\gamma_i h_i^+}} \cdot \prod_{k=i+1}^j \frac{1 - \bar{\Gamma}_k}{e^{-\gamma_k h_k} - \bar{\Gamma}_k e^{\gamma_k h_k}} \quad (2.8c)$$

$$\bar{Y}_{z'}^m = \frac{1}{\bar{Z}_{z'}^m} = Y_i^m \frac{1 - \bar{\Gamma}_i}{1 + \bar{\Gamma}_i}, \quad \bar{Y}_{z'}^m = \frac{1}{\bar{Z}_{z'}^m} = Y_i^m \frac{1 - \bar{\Gamma}_i}{1 + \bar{\Gamma}_i} \quad (2.8d)$$

The definition of Y_i^m is the same as that in (2.6); I_V^P, V_I^P can be obtained by the well-known transmission line (TL) theory.

2.2.4 SINGLE LAYER STRUCTURE

As a special case of multilayer structure, the single layer open structure for FGMSL is shown in Fig. 1.4. Both upper strip and lower finite ground plane are the perfect conductors with infinitesimal thickness. These two conductors are formed on the top and bottom surfaces of an isotropic dielectric substrate with thickness h , and permittivity ϵ_1 . In this case, electric currents will induce the EM signals as shown in

Fig. 2.3. The TE and TM modal amplitudes in the spectral DGF's can be determined by solving two sets of inhomogeneous transmission line equations (2.5).

Fig. 2.3. Equivalent electric currents in a layer medium of Finite-Ground Microstrip line. (a) Physical configuration. (b) Equivalent TL models for the transverse and vertical electric point sources, respectively.

For a three-dimensional electric point source excitation, the modal amplitudes $W^{TE}(z,0)$, $W^{TM}(z,0)$, $H^{TM}(z,0)$ are given by,

$$W^{TE}(z,0) = \begin{cases} V^{TE}(0), & z = 0 \\ V^{TE}(0) \frac{\gamma_1}{\gamma_1 ch(\gamma_1 h) + \gamma_0 sh(\gamma_1 h)}, & z = h \end{cases} \quad (2.9)$$

$$W^{TM}(z,0) = \begin{cases} V^{TM}(0), & z = 0 \\ V^{TM}(0) \frac{\epsilon_1 \gamma_0}{\epsilon_1 \gamma_0 ch(\gamma_1 h) + \gamma_1 sh(\gamma_1 h)}, & z = h \end{cases} \quad (2.10)$$

$$H^{TM}(z,0) = \begin{cases} I^{TM}(0), & z = 0 \\ I^{TM}(0) \frac{\gamma_1}{\gamma_1 ch(\gamma_1 h) + \epsilon_1 \gamma_0 sh(\gamma_1 h)}, & z = h \end{cases} \quad (2.11)$$

where

$$V^{TE}(0) = j\omega\mu_0 \frac{\gamma_0 + \gamma_1 th(\gamma_1 h)}{2\gamma_0 \gamma_1 + (\gamma_1^2 + \gamma_0^2) th(\gamma_1 h)} \quad (2.12)$$

$$V^{TM}(0) = \frac{\gamma_0 \gamma_1}{j\omega \varepsilon_0} \frac{\varepsilon_1 \gamma_0 + \gamma_1 th(\gamma_1 h)}{2\varepsilon_1 \gamma_0 \gamma_1 + (\varepsilon_1^2 \gamma_0^2 + \gamma_1^2) th(\gamma_1 h)} \quad (2.13)$$

$$I^{TM}(0) = j\omega \varepsilon_0 \varepsilon_1 \frac{\gamma_1 + \varepsilon_1 \gamma_0 th(\gamma_1 h)}{2\varepsilon_1 \gamma_0 \gamma_1 + (\varepsilon_1^2 \gamma_0^2 + \gamma_1^2) th(\gamma_1 h)} \quad (2.14)$$

$$\gamma_1^2 = k_t^2 - \omega^2 \mu_0 \varepsilon_0 \varepsilon_1 \quad (2.15)$$

$$\gamma_0^2 = k_t^2 - \omega^2 \mu_0 \varepsilon_0 \quad (2.16)$$

The other modal amplitudes, namely, $W^{TE}(z, z')$, $I_V^{TE}(z, z')$, $W^{TM}(z, z')$, $I_V^{TM}(z, z')$

and $V_I^{TM}(z, z')$ can be easily obtained from (2.5) in the same way.

Finally, we have

$$\begin{aligned} \bar{\bar{G}}_{EJ}(\bar{k}_t, z, z') = & -W^{TE}(z, z')(\hat{k}_t \times \hat{z})(\hat{k}_t \times \hat{z}) - W^{TM}(z, z')\hat{k}_t\hat{k}_t \\ & + \frac{k_t}{\omega \varepsilon_0 \varepsilon(z)} I_V^{TM}(z, z')\hat{z}\hat{k}_t - \frac{k_t}{\omega \varepsilon_0 \varepsilon(z')} V_I^{TM}(z, z')\hat{k}_t\hat{z} \\ & + \left(\frac{k_t}{\omega \varepsilon_0}\right)^2 \frac{1}{\varepsilon(z)\varepsilon(z')} II^{TM}(z, z')\hat{z}\hat{z} - \frac{\delta(z-z')}{j\omega \varepsilon_0 \varepsilon(z)} \hat{z}\hat{z} \end{aligned} \quad (2.17)$$

2.3 METHOD OF MOMENTS WITH SOC TECHNIQUE

The MoM is one of the most popular methods for solving electromagnetic field problems. It converts the integral equation into a matrix equation by using the basis (expansions) functions and weighting (testing) functions. SOC technique is an advanced de-embedding technique in the electromagnetic field theory. It can extract the equivalent circuit models of lumped elements for planar integrated circuits directly by using the admittance-type MoM algorithm [87], [88], [89]. Using this technique, the multi-port network parameters of planar circuits without receiving the parasitic effects can be effectively characterized in the impressed-source MoM platform.

2.3.1 METHOD OF MOMENTS WITH IMPRESSED VOLTAGE SOURCE

First, we consider a model of the unbounded n -port planar microwave circuit fed by microstrip feed lines, as shown in fig. 2.4. An electric-field integral equation (EFIE) can be established by enforcing the boundary condition of the tangential electric field Eq. (2.3a) on the conductor surface, as

$$\sum_{i=1}^n \vec{E}_i^{inc}(\vec{r}) + \sum_{i=1}^n \int_{line(i)} \vec{G}_i(\vec{r} / \vec{r}_s) \cdot \vec{J}_i(\vec{r}_s) dS_s + \int_{circuit} \vec{G}(\vec{r} / \vec{r}_s) \cdot \vec{J}(\vec{r}_s) dS_s = 0 \quad (2.18)$$

As described in [86], [6], the delta-gap voltage-source backed by a perfect electric wall has been proposed in the MoM algorithm to solve the complex unbounded microstrip structures. In this case, the current density $\vec{J}$ flowing along the feed lines can be expanded in terms of a linear combination of sub-sectional basis functions $\vec{J}_i(\vec{r}_s)$ and their images $\vec{J}'_i(\vec{r}_s)$ which account for the effect of the local electrical wall at the selected port location. Thus, the total current density $\vec{J}$ at the i th port in Eq. (2.18) may be written as follow:

$$\vec{J}_i(\vec{r}_s) \Rightarrow \vec{J}_i(\vec{r}_s) + \vec{J}'_i(\vec{r}_s) = \vec{J}_i(\vec{r}_s) + \vec{J}_i(\vec{r}_i - \vec{r}_s) \quad (2.19)$$

At the i th port, the incident electric field $\vec{E}_i^{inc}$ can be expressed by an impressed voltage source

Fig. 2.4. A general multi-port planar microwave circuit driven by microstrip feed lines.

$$\vec{E}_i^{inc}(\vec{r}) = \hat{n}_i \cdot V_i \cdot \delta(\vec{r} - \vec{r}_i) \quad (2.20)$$

where $\hat{n}_i$ is the longitudinal direction on the feed line, V_i is the delta-gap voltage.

Over the conductor surface of the circuit under modeling, the vector current density $\vec{J}(\vec{r}_s)$ can be expressed in terms of linear superposing of sub-domain basis functions $\vec{J}_n(x_s, y_s)$ with unknown amplitudes I_n ,

$$\vec{J}(\vec{r}_s) = \sum_{n=1}^N I_n \vec{J}_n(x_s, y_s) \quad (2.21)$$

We can transform the integral equations into their corresponding matrix equations by multiplying the weighting functions $\vec{J}_n^w(\vec{r}_s)$ into the two sides of Eq. (2.18) and integrating over the conductor surface.

With the selection of the same number (N) for basis and weighting functions, a set of N equations can be derived for numerically solving the unknown current amplitudes I_n ($n = 1, 2, \dots, N$).

Here, we choose the piecewise sinusoidal (PWS) functions in the longitudinal direction and constant (pulse) functions in transverse direction for basis and weighting functions. Fig. 2.5 shows the configuration of the basis and weighting functions at the i th port.

For the FGMSL structure illustrated in Fig. 1.4, Eq. (2.18) should be modified to consider the two strip conductors at the two different surfaces of a dielectric substrate, i.e. strip and ground surfaces.

Fig. 2.5. Configurations of the weighting and basis functions for a uniform feed line backed by vertically electric wall in the MoM. (a) Weighting functions (b) Basis functions.

$$\hat{z} \times \left[\vec{E}_1^{inc}(\vec{r}) + \iint_{strip} \bar{\bar{G}}_{EJ}(h, h) \cdot \vec{J}_1(\vec{r}_s) dS_s + \iint_{ground} \bar{\bar{G}}_{EJ}(h, 0) \cdot \vec{J}_2(\vec{r}_s) dS_s \right] = 0 \quad (2.22a)$$

$$\hat{z} \times \left[\vec{E}_2^{inc}(\vec{r}) + \iint_{strip} \bar{\bar{G}}_{EJ}(0, h) \cdot \vec{J}_1(\vec{r}_s) dS_s + \iint_{ground} \bar{\bar{G}}_{EJ}(0, 0) \cdot \vec{J}_2(\vec{r}_s) dS_s \right] = 0 \quad (2.22b)$$

By defining the basis and weighting functions as discussed above for the MSL or FGMSL circuits, we can transform the Eq. (2.18) or Eq. (2.22) into a set of equations and further derive the spectral-domain expression using the Fourier Transformation. Then we have

$$\begin{bmatrix} \langle w_1 ; f_1 \rangle & \cdots & \langle w_1 ; f_N \rangle \\ \langle w_2 ; f_1 \rangle & \cdots & \langle w_2 ; f_N \rangle \\ \vdots & & \vdots \\ \langle w_N ; f_1 \rangle & \cdots & \langle w_N ; f_N \rangle \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_N \end{bmatrix} = \begin{bmatrix} \langle w_1 ; E_1 \rangle \\ \langle w_2 ; E_2 \rangle \\ \vdots \\ \langle w_N ; E_N \rangle \end{bmatrix} \quad \text{or} \quad [Z][I] = [V] \quad (2.23)$$

where $w_i = \tilde{J}_i^w(\alpha, \beta)$, $f_i = \tilde{\bar{\bar{G}}}_{EJ}(\alpha, \beta) \cdot \tilde{J}_i(\alpha, \beta)$, $E_i = \tilde{E}_i^{inc}(\alpha, \beta)$, $z_{mn} = \langle w_m ; f_n \rangle$ and

$$V_m = \langle w_m ; E_m \rangle.$$

By numerically solving this matrix equation Eq. (2.23), we can obtain the solution for the unknown amplitudes of all the basis functions and thus calculate the equivalent currents and voltages at each port.

$$[I] = [Z]^{-1} [V] \quad (2.24)$$

In this case, the network parameters of the circuits under modeling such as Y -parameters can be derived from the calculated I_i and V_i at each port of an unbounded multi-port structure.

After the above procedure, Y -parameters of the network can be directly converted from $[Z]$ matrix in Eq. (2.24). On the other hand, S -parameters of this network can be expressed in terms of the above Y -parameters as,

$$\begin{cases} S_{11} = \frac{(1 - Z_0 Y_{11})(1 + Z_0 Y_{22}) + Y_{12} Y_{21} Z_0^2}{\Psi} \\ S_{12} = \frac{-2Y_{12} Z_0}{\Psi} \\ S_{21} = \frac{-2Y_{21} Z_0}{\Psi} \\ S_{22} = \frac{(1 + Z_0 Y_{11})(1 - Z_0 Y_{22}) + Y_{12} Y_{21} Z_0^2}{\Psi} \end{cases} \quad (2.25)$$

where $\Psi = (1 + Z_0 Y_{11})(1 + Z_0 Y_{22}) - Y_{12} Y_{21} Z_0^2$ and Z_0 is the characteristic impedance.

2.3.2 SHORT-OPEN CALIBRATION TECHNIQUE

In order to carry out the SOC procedure for calibrating out the numerical error caused by the impressed source in the above-described MoM, a planar structure may be divided into two separate parts: error boxes and a circuit box. Fig. 2.6 illustrates

Fig. 2.6. An unbounded microstrip circuit framework and its equivalent circuit model. (a) Physical layout at the i th port. (b) Equivalent circuit representation.

physical layout at the i th port of an unbounded multi-port microstrip circuit framework and its equivalent circuit model.

In Fig. 2.6(b), the impressed source causes a port discontinuity with the quasi-lumped L_p and C_p , while the 3-D definition of uniform feed line is not consistent to its 2-D definition in theory. These two aspects of error effects are put together to constitute the so-called error box as marked in Fig. 2.6(b). This SOC de-embedding procedure is utilized to characterize the parameters of this error block at each port of the circuits by defining the two calibration elements described as follows. Then we can extract the precisely realistic circuit parameters at the reference planes $\#i$ of interest from the port planes $\#i'$ which have been obtained in section 2.3.1. Here, the error box $[X_i]$ at the i th port is expressed as the ABCD matrix,

$$\begin{bmatrix} V_i' \\ I_i' \end{bmatrix} = [X_i] \begin{bmatrix} V_i \\ I_i \end{bmatrix} \quad (2.26)$$

Fig. 2.7. Physical models and circuit networks of the “Open” and “Short” elements. (a) Ideal “Open” line with even E-field excitations; (b) Ideal “Short” line with odd E-field excitations.

where V_i', I_i' are realistic circuit parameters that need to be extracted, V_i, I_i are the port voltage and current determined by Eq. (2.24).

Now, the main task is to characterize this error box with the “short” and “open” calibration elements. Compared to the traditional de-embedding technique, SOC procedure can be directly implemented in the MoM algorithm without needing to use the 2-D defined characteristic impedance. These elements can be formulated by exciting a pair of even and odd impressed electric fields at the two terminals of a uniform microstrip line, as shown in Fig. 2.7. The uniform line with the length of twice distance Li is selected to repeatedly express the electrical property of the feed line error box, where Li is defined as the length between port plane $\#i$ and reference plane $\#i'$. The symmetrical plane of this twice length feed line will become a magnetic wall in the even excitation model and an electric wall in the odd excitation model. In other words, ideal “Open-end” and “Short-end” can be constructed in the MoM at the symmetrical plane.

In this model, the uniform line with the $2L_i$ length can be regarded as a two port symmetric structure excited by two independent impressed electric fields. With Eq. (2.18), this two-port EFIE for single layer FGMSL feed line at the i th port can be easily formulated as

$$\begin{cases} \sum_{j=1}^2 \vec{E}_{ij}^{inc}(\vec{r}) + \iint_{S_{i1}} \vec{G}_{EJ}(h, h) \cdot \vec{J}_1(\vec{r}_s) dS + \iint_{S_{i2}} \vec{G}_{EJ}(h, 0) \cdot \vec{J}_2(\vec{r}_s) dS = 0 \\ \iint_{S_{i1}} \vec{G}_{EJ}(0, h) \cdot \vec{J}_1(\vec{r}_s) dS + \iint_{S_{i2}} \vec{G}_{EJ}(0, 0) \cdot \vec{J}_2(\vec{r}_s) dS = 0 \end{cases} \quad (2.27)$$

where $\vec{E}_{i1}^{inc}, \vec{E}_{i2}^{inc}$ indicate the impressed fields at the two terminals of the upper strip conductor, S_{i1} indicates the area of upper strip ($2L_i$), and S_{i2} is the area of lower strip ($2L_i$). The 3D-view of this structure is sketched in Fig. 2.8.

Based on the network theory, the voltages and currents, as in Fig. 2.7, can be separately defined for even model and odd excitation cases, that is,

Even case: V_{ie}, I_{ie} for port, $V_{ie}', I_{ie}' (=0)$ for end;

Odd case: V_{io}, I_{io} for port, $V_{io}' (=0), I_{io}'$ for end.

Then we have the following equations to obtain the matrix parameters of the two-

Fig. 2.8. 3D-View of the symmetrical structure for formulating the two SOC standards regarding to the i th FGMSL port.

port error box in terms of MoM-calculated currents at the port and short-end.

$$\begin{cases} d_i V_{io} - b_i I_{io} = 0 \\ -c_i V_{io} + a_i I_{io} = I_{io}' \\ -c_i V_{ie} + a_i I_{ie} = 0 \\ a_i d_i - b_i c_i = 1 \end{cases} \quad (2.28)$$

Through the solution of Eq. (2.28), we can obtain the $ABCD$ matrix of the i th error block in terms of normalized currents,

$$\begin{bmatrix} a_i & b_i \\ c_i & d_i \end{bmatrix} = \begin{bmatrix} \bar{I}_{io}' / (\bar{I}_{io} - \bar{I}_{ie}) & 1 / \bar{I}_{io}' \\ \bar{I}_{io}' \bar{I}_{ie} / (\bar{I}_{io} - \bar{I}_{ie}) & \bar{I}_{io} / \bar{I}_{io}' \end{bmatrix} \quad (2.29)$$

where $\bar{I}_{io} = I_{io} / V_{io}$, $\bar{I}_{io}' = I_{io}' / V_{io}$, $\bar{I}_{ie} = I_{ie} / V_{ie}$.

In order to directly apply the derived $ABCD$ matrix in Eq. (2.29), an alternative $ABCD$ matrix $[X_i]$ is given in Eq. (2.30) to express the equivalent voltage V_i' and current I_i' at the plane $\#i'$ in terms of V_i and I_i at the port $\#i$.

$$[X_i] = \begin{bmatrix} d_i & -b_i \\ -c_i & a_i \end{bmatrix} = \begin{bmatrix} \bar{I}_{io} / \bar{I}_{io}' & -1 / \bar{I}_{io}' \\ -\bar{I}_{io}' \bar{I}_{ie} / (\bar{I}_{io} - \bar{I}_{ie}) & \bar{I}_{io}' / (\bar{I}_{io} - \bar{I}_{ie}) \end{bmatrix} \quad (2.30)$$

CHAPTER 3

Full-Wave Modeling of Finite Ground Microstrip Line

3.1 INTRODUCTION

Full-wave modeling of various discontinuity elements is always a critical issue for design of the whole circuit blocks with complicated configuration. As usual, the discontinuity dimension is much smaller than the wavelength in microstrip, so that it is difficult to characterize the equivalent circuit models from the full-wave MoM simulation because of the harmful parasitic effects caused by the impressed excited fields as discussed in Chapter 2. As reviewed in [6], various microstrip discontinuities have been characterized by quasi-static analysis as well as full-wave analysis. The work here focuses on the modeling of the Finite-ground microstrip line (FGMSL) open-end circuits with different strip configurations by our MoM-SOC algorithm. After achieving numerical stability and accuracy in our SOC

scheme, we move to make extensive applications to various FGMSL circuits and discontinuities with different geometrical configurations.

FGMSL was early studied in theory as a modified microstrip line (MSL) [16] and it has been recently gaining a potential application in exploration of high-performance hybrid microwave integrated circuits using the multilayered Silicon techniques [20], [21]. Due to an additional degree of freedom in its finite-ground width, this FGMSL has several attractive electrical features over the traditional MSL, i.e. high characteristic impedance [16], low transmission loss [20] and good electromagnetic (EM) isolation [21]. So far, limited research work has been carried out to investigate the transmission behavior of a uniform FGMSL [16], [20] and coupling performance of a parallel-coupled FGMSL. However, little work has been done to characterize a variety of FGMSL circuits and discontinuities with different finite-ground dimensions. To meet this requirement, this chapter aims at quantitatively investigating the diverse finite-ground effects in the FGMSL structures through the one-port FGMSL open-end discontinuity. Subsequently, a two-port FGMSL circuit block with generalized configuration is considered and characterized. An improved transmission line electromagnetic bandgap (EBG) structure is made up by utilizing high-impedance property of an offset FGMSL.

3.2 FULL-WAVE CHARACTERISTICS OF FGMSL OPEN-END DISCONTINUITY

The microstrip line (MSL) open-end discontinuity with infinite ground plane has been extensively studied using the static and full-wave approaches [6], [79], [80], [87], [91], [92]. In this work, a FGMSL open-end discontinuity with finite-ground width and length is characterized as its unified equivalent circuit model with a

complex admittance by integrating the so-called short-open calibration (SOC) procedure [87] into the full-wave method of moments (MoM) [86], i.e. MoM-SOC technique. In the past, this technique has been successfully applied to model various MSL circuits [87], [89], [93], [94], coplanar stripline (CPS) circuits [88] and so on. In order to formulate a determinant MoM for modeling the FGMSL structure, the three possible delta-gap sources are at first introduced and described. After the numerical errors caused by these non-ideal excitations confirm to be calibrated out in our MoM-SOC, extensive results are derived to validate the extracted model parameters as compared with the available ones for the traditional MSL case and further quantitatively exhibit the actual effects of finite-ground dimensions.

3.2.1 MoM-SOC MODELING TECHNIQUE FOR ONE-PORT FGMSL CIRCUITS

Fig. 3.1(a) describes the cross section of a uniform FGMSL with unequal strip and finite-ground widths, w_1 and w_2 , at the upper and lower interfaces of a dielectric substrate. Fig. 3.1(b) depicts the top view of the FGMSL open-end discontinuity to be considered, where the lower finite-ground is longitudinally extended beyond the upper strip by ΔL . By enforcing the boundary condition that the tangential electric field on the strip and ground conductors disappears, a set of coupled electric-field integral equations (EFIE) can be readily established to numerically solve the current density over the upper and lower conductors by applying the Galerkin's technique. Similar to [86], in order to formulate a source-type MoM, a delta-gap voltage source backed by vertical electric wall needs to be introduced at the port away from the FGMSL open-end. Fig. 3.2 illustrates the schematic of the three possible source models, marked by upper-, lower- and balanced-source, in which the impressed

Fig. 3.1.1. Physical layout of the FGMSL open-end discontinuity to be modeled.

voltage is inserted into the infinitesimal gap between the upper/lower conductor and electric wall, respectively.

These source models are certainly not exactly matched to the field distribution of the dominant mode at the port since they also excite many of the other unwanted higher-order modes. But, as the feeding line length (L) is sufficiently enlarged, only the dominant mode can reach to the FGMSL open-end while the other modes eventually disappears due to their quick attenuation. For this reason, the whole FGMSL layout can be segmented into two distinct parts [87], [89]: FGMSL feeding line and open-end discontinuity. The former part includes all the parasitic effects caused by the non-ideal source models in the MoM, and it is usually called as error box [87].

By applying the SOC technique as detailed in [87]-[89], [93], [94], the electrical behavior of this error box can be independently evaluated via definition of two calibration standards, i.e. ideal FGMSL short- and open-end circuits in the

Fig. 3.2. Schematic of the three delta-gap voltage sources with different configurations for formulating the source-type MoM. (a) upper-source. (b) lower-source. (c) balanced-source.

consistent MoM platform. After this error box is calibrated out, the latter core FGMSL open-end can be effectively de-embedded or extracted, and it can be perceived as a unified equivalent circuit model with a fringing capacitance (C_{oc}) and radiation-caused conductance (G_L) [87].

3.2.2 RESULTS AND DISCUSSION

In order to assure that the above numerical error is calibrated out in our MoM-SOC, the FGMSL open-end capacitances (C_{oc}) are extracted under the three different excitation sources as described in Fig. 3.2 and they are plotted together in Fig. 3.3 as a function of the feeding line length (L).

From Fig. 3.3, it is found that regardless of different source models, the extracted C_{oc} persistently and stably converges to the same value as L increases. Of them, the balanced-source model is observed to achieve the fastest numerical convergence versus L . This can be interpreted by the fact that its excited field distribution at the

Fig. 3.3. Extracted FGMSL open-end capacitance (C_{oc}) via feeding line lengths using the MoM-SOC under three different source configurations as in Fig. 3.2.

Fig. 3.4. Comparison among extracted capacitances (C_{oc}) of an FGMSL open-end and its traditional MSL counterpart from static and full-wave algorithms.

source port is much closer to that of the dominant mode in the FGMSL rather than the two others. As a result, this balanced model is utilized for numerical implementation.

To evidently validate our extracted results, the FGMSL open-end discontinuity with equal upper strip and low finite-ground widths, i.e. $W_1=W_2=W$, is modeled

here and the relevant normalized capacitance (C_{oc}/W) is plotted in Fig. 3.4 together with those of its traditional MSL counterpart with infinite ground plane.

Due to its geometrical symmetry with respect to the horizontal central plane of the substrate, half an FGMSL structure is exactly same as its corresponding MSL one with the half thickness. From Fig. 3.4, the extracted FGMSL C_{oc}/W is found about 25.2pF/m for $h=1.27\text{mm}$ and thus its double value represents that of the MSL open-end with half thickness, i.e. $h=0.635\text{mm}$. As a result, the latter one can be directly compared with those obtained from the static approach [6] and the MoM-SOC [87] for the case of infinite ground plane and it is found in good agreement with them as illustrated in Fig. 3.4.

Now, we move to investigate the effects of finite-ground dimension on the FGMSL open-end discontinuity via extracted equivalent complex admittance ($Y_L=G_L+jB_L$). Fig. 3.5(a) illustrates these two parameters versus different ground widths (w_2) for a FGMSL open-end with the fixed strip width ($W_1=0.6\text{mm}$) and extended length ($\Delta L=0\text{mm}$). Herein, B_L rises up as a linear function of the frequency for all the listed cases, thereby indicating the static performance of a fringing capacitance (C_{oc}) as discussed above. As W_2 is widened, its frequency-related slope or C_{oc} increases quickly first and then slowly.

Meanwhile, G_L is found negligibly small over the frequency range of 2.0 to 5.0 GHz. Next, both strip and ground widths are kept unchanged, i.e. $W_1=0.6\text{mm}$ and $W_2=1.8\text{mm}$, while the excess finite-ground length (ΔL) beyond the open-end location is gradually extended. The relevant results are plotted in Fig. 3.5(b) and they demonstrate that B_L goes up with ΔL in a decelerated manner and eventually

(a)

(b)

Fig. 3.5. Extracted admittance $Y_L = G_L + jB_L$ of FGMSL open-end discontinuities with different dimensions. (a) Finite-ground widths (W_2). (b) Finite-ground lengths (ΔL).

converges to the unchanged straight line as ΔL is enlarged beyond 8.0 mm. As observed in Fig. 3.5(b), G_L is still kept close to zero.

3.2.3 CONCLUSION

In this work, the MoM-SOC technique is extended to full-wave characterization of a generalized finite-ground microstrip line (FGMSL) open-end discontinuity with

different configurations and further parametric extraction of its unified equivalent circuit model. After the error term in the determinant MoM is calibrated out and the extracted model parameter is confirmed with respect to the published ones, extensive results are provided to investigate its dynamic circuit model parameters without any hypothesis in theory. It is exhibited here that the FGMSL open-end discontinuity can still be modeled as a simple static fringing capacitance and its value increases as the finite-ground dimension is enlarged.

3.3 OFFSET FGMSL WITH USING HIGH-IMPEDANCE PROPERTY FOR EBG ENHANCEMENT

During the past decades, periodic structures have been raising the continuous interest in the development of RF and microwave integrated circuits because of their slow-wave and frequency filtering properties. With the series-inductive or shunt-capacitive loading in periodic intervals, non-uniform planar transmission lines exhibit the stopband characteristics in a certain range [12]. This periodic structure with stopband was recently referred to as the photonic bandgap (PBG) [94] or electromagnetic bandgap (EBG) [95]. In the past years, various EBG structures have been developed in exploring advanced circuit blocks [94], suppressing unwanted modes in shielded planar circuits [95], eliminating harmful noises in high-speed printed circuit boards [96], and so on.

On the other hand, planar EBG structures have been more comprehensively investigated in terms of the two effective per-unit-length transmission parameters of periodic structures with infinite extension [97]-[99]. In particular, the non-zero attenuation constant and imaginary characteristic impedance in the stopband have been reported in [99] to display the complete guided-wave characteristics of

Fig. 3.6. 3-D geometry of periodic electromagnetic bandgap (EBG) finite-ground microstrip line. (a) Unit-cell. (b) N-cell.

microstrip EBG. In this part work, our effort is focused on achieving the enhanced EBG performance using the high-impedance property of offset finite-ground microstrip line (FGMSL). Fig. 3.6 depicts the three-dimensional (3-D) geometry of the proposed EBG structure based on the periodic FGMSL. Following up with the one-port modeling technique in Section 3.2, a two-port FGMSL circuit block with generalized configuration is considered and characterized by using a self-calibrated MoM. The effective propagation constant and characteristic impedance of this EBG structure are extracted to demonstrate the enhanced guided-wave bandgap behavior with widened stopband. The magnitude and phase quantities of transmission coefficients for a five-cell EBG circuit are then displayed under ideal impedance matching at the two ports. This EBG circuit with the two 50Ω feed lines is further constructed and characterized to demonstrate the enhanced EBG performance existed in an actual planar EBG circuit.

Fig. 3.7. Full-wave modeling and numerical extraction of a generalized two-port FGMSL circuit or discontinuity in the self-calibrated MoM platform. (a) Physical schematic. (b) Equivalent cascaded topology.

3.3.1 MODELING OF TWO-PORT FGMSL

Let's start with a brief description on the modeling of a generalized two-port FGMSL structure using the self-calibrated MoM to deal with the infinite-ground MSL discontinuities. It is applied in Section 3.2 for extraction of an FGMSL open-end fringing capacitance. Fig. 3.7 shows the physical layout and equivalent-circuit topology for modeling and de-embedding of such a two-port FGMSL circuit or discontinuity block by executing the SOC procedure in the source-type MoM platform.

In order to formulate a determinant MoM, two impressed delta-gap voltage sources V_{P1} and V_{P2} backed by vertical electric wall are introduced at the two ends or terminals on the left- and right-side of the FGMSL feed lines. The two source ports P_1 and P_2 should be readily selected far away from the FGMSL circuit at the center such that only the dominant mode can reach the two reference planes R_1 and R_2 with the disappearance of all other modes. By doing so, the current densities over the upper and lower strip/ground conductors can be derived through the

numerical solution of an electric-field integral equation (EFIE) according to the well-known Galerkin's method. As a result, the two-port network parameters at the two source ports, P_1 and P_2 , can be initially obtained in terms of the solved currents, I_{P1} and I_{P2} , and the impressed voltages, V_{P1} and V_{P2} [87], [89].

Next, the overall FGMSL layout is classified into three distinct sections, i.e. the two feed lines and the core FGMSL circuit at the center, as illustrated in Fig. 3.7(a). Fig. 3.7(b) indicates the relevant equivalent cascaded-network topology. The two error boxes, $[X_1]$ and $[X_2]$, represent the overall dynamic effects from the source ports to the reference planes of the two feed lines, including the effects of the non-ideal impressed sources at the ports. Each of them can be separately characterized with the use of a pair of FGMSL short and open standard elements, which are formed and defined in the same MoM. Upon executing such a SOC procedure, the equivalent circuit model at the center can be simply deembedded by removing the two error boxes out from an overall network on a basis of a cascaded transmission line theorem. In this way, the core circuit model is directly presented as an ABCD matrix. Its four known elements can be utilized to define the three-dimensional (3-D) characteristic impedance provided that the core circuit is a uniform FGMSL with finite length [102]. Moreover, it also can be used to derive the impedance- or admittance-matrix of various FGMSL circuit structures [87], [89].

3.3.2 EBG GEOMETRY AND PARAMETRIC EXTRACTION

As illustrated in Fig. 3.6, the proposed EBG structure is constructed on the basis of the FGMSL [100] with periodic configuration on the upper and lower finite-width strip conductors. Looking at the schematic of a single EBG unit cell in Fig. 3.6(a),

the two lower and upper strip conductors in the central part are simultaneously narrowed and then transversely separated so as to make up an offset FGMSL with high characteristic impedance. Meanwhile, the two strips at the two sides are equally widened to form an FGMSL with low characteristic impedance. In this way, the former one can be perceived as an equivalent series-inductive element, while the latter is equivalent to a shunt-capacitive element. It can be intuitively understood that the equivalent series-inductive quantity in the offset FGMSL can be largely enhanced by separating the two strip conductors away from each other.

In the modeling, periodic FGMSL with the cell number of N is linked with the two uniform FGMSL feed lines and its performance can be in general characterized using the self-calibrated method of moments [100]. As pointed out in [99], this N -cell periodic FGMSL or EBG structure can be considered as an equivalent uniform transmission line section, with the length of $L=NT$, that may have the frequency-dispersive, full-attenuation or full-transmission performances. To investigate the fundamental guided-wave characteristics in this infinite-extended EBG, the two effective per-unit-length parameters of this artificial transmission line, i.e. complex propagation constant ($\gamma = \alpha + j\beta$) and complex characteristic impedance ($Z_0 = \text{Re}(Z_0) + j \text{Im}(Z_0)$), need to be derived as discussed in [97]-[99]. They can be explicitly expressed in terms of the four calculated elements, A , B , C and D , of the two-port ABCD matrix defined at the interfaces between the periodic and uniform FGMSL [99].

$$Z_0 = \sqrt{\frac{B}{C}} \quad \text{and} \quad \cosh(\gamma L) = \frac{A + D}{2} \quad (3.1)$$

3.3.3 RESULTS AND DISCUSSION

Fig. 3.8(a) and Fig. 3.8(b) depict the two extracted complex transmission parameters of the periodic FGMSL with different strip widths of $W_1=0.6$ and 1.2 mm, respectively. In these two cases, two narrow strip conductors of the offset FGMSL are always put at the extremely outer edges of the wide strip conductors as shown in Fig. 3.6(a) and Fig. 3.6(b). In the low frequency range, the normalized

(a)

(b)

Fig. 3.8. Extracted per-unit-length complex transmission parameters of infinite-length periodic FGMSL. (a) Normalized propagation constants. (b) Characteristic impedances.

propagation constant (γ/k_0) has only imaginary part and this normalized phase constant (β/k_0) is significantly raised as W_1 is widened from 0.6 to 1.2 mm. It is because that the widened W_1 leads to enlargement of the coupling between the lower and upper strip conductors in the symmetrical FGMSL while the increased offset distance (W_1-W_2) tremendously reduces this coupling in the offset FGMSL. As the frequency increases, the guided-wave becomes slow in propagation region and then moves itself into the attenuation region, namely, bandgap in [94]-[96], with the non-zero imaginary impedance and attenuation constant, as also comprehended in [99], [101]. Looking at the two sets of graphs together, we can find out that the bandgap has really gained large enhancement from 41% to 81% in fractional bandwidth. Beyond this bandgap, the guided-wave enters into its next passband region [99], [101].

Let's now consider that the guided-wave propagates across the truncated finite-cell periodic FGMSL that is assumed to be ideally matched with the source and load impedances. This arrangement is made at first to illustrate the propagation performance in this EBG structure without receiving any unexpected mismatching effects at two interfaces. On the basis of the cascaded transmission line theorem, the transmission coefficient (S_{21}) of this periodic EBG transmission line with the length of $L=NT$ can be simply expressed as below, where the impedance (Z_0) disappears due to ideal matching.

$$S_{21} = e^{-\alpha L - j\beta L} \quad (3.2)$$

Furthermore, the magnitude ($|S_{21}|$) and phase (Φ_{21}) can be separately derived in an explicit format.

$$|S_{21}| = e^{-\alpha NT} \quad \text{and} \quad \Phi_{21} = -\beta NT \quad (3.3)$$

Fig. 3.9(a) and Fig. 3.9(b) show the two sets of these parameters for the above-discussed periodic EBG structures with five cells, i.e. $N=5$. In the low and high passbands, the magnitude ($|S_{21}|$) is observed in Fig. 3.9(a) as the unity constant in both cases since the attenuation constant (α) is consistently equal to zero. In the bandgap region between them, this parameter falls down and approaches the

(a)

(b)

Fig. 3.9. Magnitude and phase of transmission coefficients of five-cell periodic FGMSL EBG structures with perfect impedance matching at two terminals. (a) Magnitude: $|S_{21}|$. (b) Phase: Φ_{21} .

minimum as the frequency is moved to the center of this stopband. We can see here that the bandgap in $W_2=1.2$ mm is much wider in bandwidth and deeper in attenuation than that in $W_1=0.6$ mm. This phenomenon is fully controlled by the frequency region and magnitude of non-zero α as plotted in Fig. 3.8(a). Later on, it will be exhibited that the guided-wave bandgap characteristics with non-zero α actually determines the stopband width and insertion loss of the finite-length EBG circuit with 50Ω feed lines. The phase (Φ_{21}) of these two EBG structures is plotted in Fig. 3.9(b). Regardless, this parameter with negative quantity goes down in the lowpass band, becomes a constant of -5π in the whole bandgap, and then moves down again in the upper passband.

Finally, the above five-cell periodic EBG structures are connected with the actual 50Ω feed line at the two truncated terminals. Similarly, the relevant scattering parameters can be calculated in terms of the 50Ω source/load impedance, the above-extracted Z_0 and γ as well as the finite-length L . Fig. 3.10(a) and Fig. 3.10(b) show the calculated insertion loss $|S_{21}|$ and return loss $|S_{11}|$ based on the simple transmission line theorem together with those obtained by directly simulating the whole layouts of these EBG circuits using the Agilent Momentum, as shown in Fig. 3.11. Look at Fig. 3.9(a) and Fig. 3.10(a), the bandstop or bandgap in both two cases appears in the almost same frequency region. Several ripples observed in low and high passbands in Fig. 3.10(a) are mainly contributed by the multiple reflections between the two interfaces along the truncated EBG structure. However, the results in Fig. 3.10 show us again that the EBG performance has been really enhanced as W_1 is widened from 0.6 to 1.2 mm.

Fig. 3.10. Scattering parameters of five-cell periodic FGMSL EBG circuits with 50Ω external feed lines. (a) Transmission coefficient: $|S_{21}|$. (b) Reflection coefficient: $|S_{11}|$.

3.3.4 CONCLUSION

In this part work, a novel periodic electromagnetic bandgap structure is presented, characterized and implemented. Using the high-impedance property of the offset finite-ground microstrip line, the bandwidth and attenuation depth of the concerned guided-wave bandgap are quantitatively confirmed to gain significant enhancement. The two effective per-unit-length parameters are derived to display the fundamental guided-wave characteristics of the FGMSL EBG with infinite-extended length [97]-

Fig. 3.11. 3-D geometry of the whole layout of FGMSL EBG circuit with 50Ω feed lines.

[99]. The scattering parameters of the finite-cell EBG circuits with varied feed-line impedances are further investigated to exhibit the bandstop behaviours as a two-port filtering circuit [94]-[96].

CHAPTER 4

Stopband-Enhanced and Size-Miniaturized LPFs Using Offset FGMSL

4.1 INTRODUCTION

Conventional transmission line lowpass filters (LPFs), such as stepped-impedance, open-stub and semi-lumped element filters, have been widely used in RF and microwave applications [103]. Due to the frequency-distributed behavior of finite-extended transmission lines, these filters always suffer from poor stopband performance beyond the cutoff frequency [103], [104]. Particularly, the strong surface-to-surface coupling between the narrow strip and wide ground in the conventional microstrip line (MSL) obstructs one from raising the concerned characteristic impedance sufficiently. For this reason, the MSL is formed with narrowed strip width and extended line length to realize the equivalent high series inductance as usually used in the lowpass filter design. In this way, the first harmonic passband is shifted downward to construct a narrow stopband. To

overcome these problems, various approaches have been developed, such as raising the characteristic impedance effectively by forming a backside aperture underneath the strip conductor [105] and utilizing a defected ground structure [106]. In addition, a compact elliptic-function LPF with wide stopband [107] is constructed using the microstrip stepped-impedance hairpin resonator. Moreover, a wide stopband can be achieved by combining the microstrip open stubs and coupled slots on the ground plane [108].

On the other hand, finite-ground microstrip line (FGMSL) has been found a variety of potential applications in the design of hybrid and multilayered microwave integrated circuits because of its distinct features, i.e. high characteristic impedance [16], low transmission loss [20] and good electromagnetic isolation [21]. In contrast to the conventional MSL, the finite-width ground plane of the FGMSL provides us with an additional degree of freedom in practical design. This will be exhibited in this work through the design of an FGMSL lowpass filter as shown in Fig. 4.1. The

Fig. 4.1. Geometry of the proposed FGMSL LPF with the upper strip conductor, middle substrate layer and lower ground plane.

FGMSL with offset strip and ground in horizon has more advantageous capacity in raising the characteristic impedance effectively than the approaches in [105], [106], as inquired in building up a harmonic-suppressed LPF.

Following up with the work in Chapter 3, characteristic impedances of the offset FGMSL are numerically extracted to quantitatively demonstrate that its quantity rises up to a great extent as the offset distance between the upper strip and lower ground with narrow widths gradually increases. Next, the two finite-length FGMSLs with high and low impedances are built up and serve as equivalent quasi-lumped series-inductive and shunt-capacitive elements, respectively. Their equivalent network parameters are numerically derived to provide us an efficient capability in designing the whole LPF in the format of a cascaded network topology. Finally, the FGMSL LPF block with enhanced performance is designed optimally. The predicted S-parameters are verified using commercial simulator and measurement taken from the fabricated LPFs.

4.2 EXTRACTED PARAMETERS OF FGMSL ELEMENTS

In this section, the two-port modeling technique of FGMSL in Section 3.3.1 is deployed to extract the characteristic impedance of a uniform FGMSL line with finite length, and the network parameters of two FGMSL quasi-lumped elements. These results will be used to carry out the physical explanation and network-based optimization of the proposed FGMSL LPF.

4.2.1 OFFSET FGMSL WITH HIGH IMPEDANCE

Fig. 4.2(a) and 4.2(b) describe the cross section of a conventional MSL and offset FGMSL, respectively. By narrowing the strip width of the former one, its characteristic impedance may be raised to a certain degree, but usually is not high enough in the filter design. On the contrary, the latter one, namely, offset FGMSL, is constructed by forming a finite-ground conductor with the same width (W) as that of the strip and further separating the upper/lower strip conductors in horizon. Intuitively, it can be understood that as the offset distance (p) increases, the coupling between the upper and lower conductors becomes weaker. Thus, the characteristic impedance of this offset FGMSL is supposed to increase to a great extent if a large offset distance (p) is selected.

Fig. 4.2. Cross-section view of the two distinctive microstrip lines (MSLs). (a) Conventional MSL. (b) Offset FGMSL.

Fig. 4.3(a) and 4.3(b) are the extracted characteristic impedance (Z_0) and normalized phase constant (β/k_0) of the offset FGMSL respectively versus offset distance (p) with verification using Agilent Momentum. In contrast to $Z_0 = 74.3\Omega$ of an infinite-ground MSL, Z_0 of the FGMSL with $p = 0\text{mm}$ (named as the double-side paralleled-stripline in [109]) becomes 121.0Ω , implying that the finite-width ground itself can increase the impedance to a certain extent. As the offset p is enlarged from

Fig. 4.3. Characteristic impedance and normalized phase constant of the offset FGMSL against those of the infinite-ground MSL with the same strip width ($w=0.2\text{mm}$). (a) Characteristic impedance. (b) Phase constant.

0 to 2.0mm, this Z_0 further rises up and reaches to 208.4 Ω . At $p=2.0\text{mm}$, the ratio of the two impedances is approximately three as shown in Fig. 4.3(a), thereby, exhibiting the significant enhancement of this impedance. The relevant β/k_0 seems to drop together with p from 2.64 to 2.13 due to the electromagnetic field dispersed in both the substrate and air regions. As a result, such an offset FGMSL with high

impedance can be used to form the improved MSL series inductive element with enhanced inductance.

4.2.2 FGMSL SERIES INDUCTIVE ELEMENT

Fig. 4.4(a) shows the geometry of an FGMSL series inductive element with the length(t_L) and strip/ground width(W_L). In order to reflect the actual situation of its application in filter design, this element is driven with the two low-impedance FGMSL with equal strip/ground widths. As illustrated in Fig. 4.4(b), its relevant

Fig. 4.4. Geometry and equivalent-circuit model of FGMSL series inductive elements. (a) Layout view. (b) Equivalent-circuit model.

equivalent π -type circuit model consists of a series reactance(X_g) and two shunt susceptances(B_p). These quantities can be numerically derived via the above-described modeling technique by selecting the two step interfaces, R_1 and R_2 , as the two reference planes in the SOC deembedding.

Fig. 4.5(a) and 4.5(b) describe the extracted frequency-related network parameters under different lengths (t_L) for the FGMSL series inductive elements in Fig. 4.4.

Fig. 4.5. Frequency-dependent π -network parameters of the FGMSL series inductive element. (a) Series reactance. (b) Shunt susceptance.

Over the frequency range of 0.5 to 12.0GHz, the series reactance (X_g) increases linearly at low frequencies, reaches a maximum, and then decreases with frequency. The frequency of maximum reactance corresponds to where the length (t_L) is one-half of a wavelength. As t_L increases from 1.4 to 3.4mm, the quasi-lumped series inductance(L_g) gets to be enhanced at the low frequency range. However, it leads to the resonant frequency being shifted down, thus lowering the first harmonic

resonance in the LPF. Meanwhile, the shunt susceptance (B_p) increases as a quasi-linear and then a tangent function of frequency due to the extended electrical length of the FGMSL as illustrated in Fig. 4.5(b). Yet, its actual quantity is still smaller compared with X_g especially at the low frequency range.

4.2.3 FGMSL SHUNT CAPACITIVE ELEMENT

Next, an FGMSL shunt capacitive element is constructed by equally widening the upper/lower conductor widths (W_C), as shown at the central part of Fig. 4.6(a). This element is fed by the two high-impedance offset FGMSL. Fig. 4.6(b) is its equivalent T-type circuit model with a shunt susceptance (B_g) and two series reactances (X_p). As width (W_C) is widened, B_g and the slope of B_g with frequency increase with increasing t_c . That is same as the case of low-impedance MSL. Similarly, the relevant T-network parameters can be extracted with reference to the two step interfaces, namely, R_1 and R_2 , as in Fig. 4.6(a).

Fig. 4.6. Geometry and equivalent-circuit model for FGMSL shunt capacitive elements. (a) Layout view. (b) Equivalent-circuit model.

Fig. 4.7(a) and 4.7(b) plot the extracted frequency-dependent network parameters under three different finite lengths (t_c) for an FGMSL shunt capacitive element. In analogy to the results illustrated in Fig. 4.5, the shunt susceptance (B_g) increases at low frequencies and then decreases as the frequency increases further. Meanwhile, the shunt reactance (X_p) increases monotonically as a quasi-linear and then tangent function. Furthermore, as t_c is extended, B_g gains uplift at the low frequency range, but it appears to be saturated and then starts to drop with the frequency

(a)

(b)

Fig. 4.7. Frequency-dependent T-network parameters of the FGMSL shunt capacitive elements. (a) Shunt susceptance. (b) Series reactance.

around the half-wave resonance point.

4.3 NOVEL FGMSL LOWPASS FILTERS

In this section, the FGMSL inductive and capacitive elements discussed above are utilized to construct, design and implement a new class of lowpass filters with widened and deepened stopband. With cutoff frequency fixed at 2.7GHz, the three 5-stage lowpass filters are designed using the MSL and FGMSL elements in order to comparatively demonstrate the attractive performances of the FGMSL-constituted LPF. Fig. 4.8(a) shows the schematic of the conventional MSL LPF, whereas the two proposed FGMSL LPFs are illustrated in Fig. 4.8(b) and 4.8(c), respectively.

Fig. 4.8. Schematics of the three LPFs. (a) conventional MSL LPF: type-A; (b) FGMSL LPF: type-B; (c) FGMSL LPF: type-C.

Fig. 4.9(a) and 4.9(b) show the basic lumped LC circuit of a lowpass filter and the equivalent lumped/distributed-element circuit based on FGMSL series inductive elements, as depicted in Fig. 4.5. Due to the increase in characteristic impedance of

Fig. 4.9. Lumped-element lowpass filters. (a) Basic lumped LC circuit. (b) Equivalent lumped/distributed-element circuit based on FGMSL series inductive elements with $W_L=0.2\text{mm}$, $t_L=2.4\text{mm}$ and $p=1.2\text{mm}$.

TABLE 4.1
DIMENSIONS OF THREE DESIGNED LPFS

TYPE	l_1	l_2	l_3	W_1	W_2	W_3
A	2.33	5.57	4.02	3.00	0.20	3.00
B	2.74	2.40	4.73	3.00	0.20	3.00
C	2.74	2.40	3.13	3.00	0.20	5.00

Unit: mm. $l_1 = l_5$, $l_2 = l_4$, $W_1 = W_5$, $W_2 = W_4$. Offset distance: $p=1.2\text{mm}$ (type-B and type-C).

the offset FGMSL, the lengths (l_2 and l_4) of the series inductive elements can be significantly shortened from 5.57 (type-A LPF) to 2.40mm (type-B and type-C LPFs). Under the same specifications in the low passband, Table 4.1 illustrates the lengths and widths of all the five strip sections after optimization on a basis of cascaded circuit networks, as shown in Fig. 4.9(b). The overall lengths of these three LPFs are derived as 19.82, 15.01 and 13.41mm, resulting in size reduction about 24% and 33% respectively, provided the high-impedance FGMSL section with short length is employed.

Fig. 4.10 plots the predicted insertion losses of these three optimized LPFs together with those calculated from the basic lumped-element LC circuit in Fig.

Fig. 4.10. Predicted insertion loss of the three LPFs shown in Fig. 4.9.

4.9(a). For the type-A LPF in Fig. 4.8(a), the lengths (l_2 and l_4) of the two narrow strip-width sections are largely extended in order to achieve the specified high inductance at the low frequency range. Thus, these lengthened sections lead to produce the first harmonic passband at 8.3GHz due to the half-wavelength resonance. For the type-B LPF in Fig. 4.8(b), it can be observed that the above harmonic is fully suppressed as l_2 and l_4 are reduced from 5.57 to 2.40mm since the shortened FGMSL with enhanced impedance is able to achieve the same series inductance at the low frequency range as that realized in the type-A LPF. However, there still exists the second harmonic passband at around 9.8GHz. This is due to the half-wavelength resonance that happens at the low-impedance FGMSL section with the length (l_3).

To further suppress this latter harmonic, the central FGMSL is properly widened in width and shortened in length under the unchanged equivalent shunt-capacitance. As can be seen in Table 4.1, the length (l_3) is shortened from 4.73 to 3.13mm. Thus,

this latter harmonic is moved upward to a great degree beyond 12.0GHz, thereby making a widened and deepened stopband as shown in Fig. 4.10.

To assure the attractive performance of the above-demonstrated FGMSL LPF, the whole layouts of the type-B and type-C filters are simulated again with Ansoft HFSS simulator at the cost of intensive CPU time. Furthermore, the relevant circuit samples are fabricated for experimental validation. To facilitate the good impedance matching with the two SMA connectors, the ground plane at the two sides is largely widened to formulate the conventional MSL feed lines. In experiment, the fabricated circuits are directly suspended or mounted in the universal substrate test fixture (WK-3000: supplied by the Inter-Continental Microwave, as shown in Fig. 4.11). In this way, the two feed lines with widened ground planes are electrically connected to the “earth” of this test fixture.

Fig. 4.11. Photograph of the universal substrate test fixture.

Fig. 4.12. Photographs of a fabricated FGMSL LPF: type (B). (a) Top view. (b) Bottom view.

Fig. 4.13. Predicted and measured S-parameters of the FGMSL LPF that is shown in Fig. 4.12.

Fig. 4.12(a) and 4.12(b) show the top/bottom-view photographs of the type-B filter. Fig. 4.13 illustrates the three sets of the predicted and measured S-parameters with a good agreement between them.

Furthermore, the type-C filter is fabricated and its top/bottom-view photographs are shown in Fig. 4.14(a) and 4.14(b). The measured results are plotted in Fig. 4.15 together with those predicted, thus ensuring the attractive stopband-enhanced performance. As can be observed from the measured results in Fig. 4.13 and Fig. 4.15, some ripples or peaks in the stopband with $|S_{21}| \leq -20\text{dB}$ is mainly caused by the

Fig. 4.14. Photographs of a fabricated FGMSL LPF: type (C). (a) Top view. (b) Bottom view.

Fig. 4.15. Predicted and measured S-parameters of the FGMSL LPF that is shown in Fig. 4.14.

imperfect impedance mismatch at the coax-to-microstrip transition in experimental implementation.

To further investigate the performance in the low passband, the three sets of the measured results with the frequency scale from 1 to 3GHz are plotted in Fig. 4.16 in conjunction with the conventional case (type-A) and the two proposed filters (type-B and type-C), respectively. Within the low passband, the type-A filter achieves 0.3dB in insertion loss and 13dB in return losses. Meanwhile, the type-B and type-C filters have the insertion losses less than 0.7dB and 1.4dB, and the return losses less than 12.7 dB and 10.3dB.

Fig. 4.16. Comparison of measured results for the three fabricated lowpass filters in Fig. 4.8.

4.4 CONCLUSION

In this section, the offset finite-ground MSL with narrow strip/ground width is studied to show its enhanced characteristic impedance and applied as an improved series inductive element for harmonic-suppressed lowpass filter design. By forming the shortened offset-FGMSL with high impedance, a new class of lowpass filters with miniaturized-size and enhanced-stopband are then proposed, designed and realized. Their advantageous filter performance is quantitatively exhibited and discussed in comparison with the conventional MSL LPF. In short, these FGMSL filters are designed optimally on a basis of a cascaded circuit network and the predicted performance is confirmed by the HFSS-based full-wave simulation and by measurement of the fabricated FGMSL LPF samples.

CHAPTER 5

Guided-Wave Characteristics of Periodically Nonuniform Coupled Microstrip Lines

5.1 INTRODUCTION

Coupled microstrip lines (CMLs) with two strip conductors placed parallel in close proximity with each other, as illustrated in Fig. 5.1(a), have been extensively studied and utilized as basic circuit elements for directional couplers, bandpass filters and so on [6]. Due to the inhomogeneous dielectric medium, the two dominant modes, i.e. even- and odd-mode, exist in this CML with different velocities of propagation. This nonsynchronous feature deteriorates the performances of microstrip circuits using the uniform CML, such as low directivity in directional coupler [46] and spurious harmonic passband in bandpass filter [47]. Extensive work has been done so far towards the equalization of propagating velocities for these two modes at certain frequencies by forming periodically-varied coupled-slot configurations, e.g., wiggly line [46], [110], zigzag line [111],

Fig. 5.1. Geometry of the uniform and various periodically nonuniform coupled microstrip lines to be considered. (a) Uniform CML. (b) PNCML (A). (c) PNCML (B). (d) PNCML (C).

corrugated line [47], [112]. On the other hand, these periodic structures are strongly expected to miniaturize the overall size of various microwave circuits by using their slow-wave behavior at the cost of extra transmission loss with lowered Q-factor.

In order to investigate in depth these periodically nonuniform coupled microstrip lines (PNCML) for circuit design, it is commonly recognized as the most critical issue to characterize the fundamental per-unit-length transmission parameters of even- and odd-mode, i.e. phase constants and characteristic impedances. Under the static assumption that each periodic unit is extremely shorter than the wavelength,

propagating velocities and characteristic impedances of the coupled striplines were approximately obtained via cascaded lumped-capacitances and inductances [110]. According to the Floquet's theorem, the two-dimensional (2-D) spectral-domain method of moments (MoM) was developed without including complicated 3-D effects to calculate the only phase constant of the two modes in PNCML [111]. Recently, the finite-difference time-domain (FDTD) method was employed to obtain the even- and odd-mode phase constants of PNCML [112]. But, to our knowledge, no research work to date has been reported to directly model the characteristic impedances of these two propagating modes in any PNCML structure with the 3-D full-wave approaches [112]. Following the description in [12], these impedances are in fact referred as to the characteristic impedances of the forward or backward Bloch waves and they are usually defined via ABCD-matrix parameters or terminal currents/voltages of a single periodical unit cell.

In this chapter, the finite-cell PNCML driven by the uniform CML lines at its two sides is characterized in the full-wave MoM platform and further the effective per-unit-wavelength transmission parameters of even- and odd-mode propagating along the PNCML are extracted via short-open calibration (SOC) technique [87]. In the past, this hybrid technique has been utilized in accurate characterization of microstrip circuits with single and multiple propagating modes at each single physical port [87], [93], [113]. Moreover, this technique has been very recently applied to extract the characteristic impedances and propagation constants of periodically inductive-loaded coplanar waveguide (CPW) [101], microstrip line [99] as well as even- and odd-mode of the uniform CPW [114]. This MoM-SOC is extended here to characterize various PNCML structures, as shown in Fig. 5.1(b-d),

targeting the numerical extraction of both characteristic impedances and phase constants for the two dominant modes. Extensive results are obtained to demonstrate the frequency- and periodicity-dependent guided-wave characteristics of these PNCML via impedance and phase constant. To validate our derived per-unit-length parameters, S-parameters of a 3-cell PNCML circuit are calculated via transmission line theorem and are then compared with those from the Agilent Momentum simulator [115] for both even- and odd-mode cases.

5.2 MoM-SOC CHARACTERIZATION OF PNCML

Fig. 5.2(a) describes the physical layout arranged for MoM-SOC modeling of a PNCML structure with finite cells (N), which is driven by the two uniform CML feed lines. In order to formulate a determinant admittance-type MoM scheme, the

Fig. 5.2. Physical layouts of the N -cell PNCML under de-embedding and the two sets of short-open-calibration (SOC) standards. (a) PNCML under de-embedding. (b) Short-circuit standards. (c) Open-circuit standards.

two pairs of delta-gap sources backed by vertical electric wall (E.W.) are

simultaneously introduced at the two strip terminals of the left- and right-side CML feed line, i.e. V_{1a} , V_{1b} at port 1 and V_{2a} , V_{2b} at port 2. As long as the two feed lines are selected electrically long, only the dominant even- and odd-mode can reach to the central PNCML section under consideration while the other excited higher-order modes at each port quickly attenuate and disappear at the reference planes, P_1 and P_2 , due to their frequency regions of operation below cutoff frequencies [87].

Following our previous work in modeling the even- and odd-mode guided-wave characteristics of the uniform coplanar waveguide [114], the even- and odd-mode in the uniform CML and PNCML can also be separately excited in the full-wave MoM platform. In this way, the even or odd modes can be generated by simultaneously enforcing that $V_{1a} = V_{1b} = V_1^e$ and $V_{2a} = V_{2b} = V_2^e$ or $V_{1a} = -V_{1b} = V_1^o/2$ and $V_{2a} = -V_{2b} = V_2^o/2$. Herein, V_1^e and V_2^e are the port voltages at port 1 and 2 under even excitation, whereas V_1^o and V_2^o are the counterpart voltages under odd excitation. As detailed in [87], the port currents at each port for both even- and odd-mode can be explicitly solved as the solution of a MoM matrix equation with voltage sources via numerical discretization of current densities over the strip conductors.

In order to de-embed the network parameters of the core PNCML at center in Fig. 5.2(a), the SOC technique [87] is employed here relying on the even- and odd-mode SOC calibration standards defined in the MoM. Fig. 5.2(b) and Fig. 5.2(c) describe the physical formulation of these two sets of SOC standards in the MoM format. Regardless of varied standards, the total length of each PNCML section in Fig. 5.2(b) and Fig. 5.2(c) must be at first selected twice the port-to-reference distance of

the considered PNCML feed line in Fig. 5.2(a). By enforcing that $V_1^e = -V_2^e$ or $V_1^o = -V_2^o$, the central plane of the uniform PNCML of $2L_f$ indicates the perfect E.W. as marked in Fig. 5.2(b), thereby constructing the ideal even- or odd-mode short-circuit standard. Similarly, the even- or odd-mode open-circuit standards can be established with central magnetic wall (M.W.) by exciting the two ports in an anti-parallel way, i.e. $V_1^e = +V_2^e$ or $V_1^o = +V_2^o$, as demonstrated in Fig. 5.2(c).

As these SOC standards are characterized in the MoM, the ABCD matrix of the two-port PNCML section with finite length of $L_p = NT$ can be numerically de-embedded by calibrating out those of feed line sections or error boxes at the two sides as discussed in [87, [99], [101], [114]. Similarly to [101], [99], the core PNCML is then equivalently perceived as a uniform CML section with effective phase constant (β_e or β_o) and characteristic impedance (Z_{0e} or Z_{0o}), that are frequency-dispersive and periodicity-dependent, in conjunction with even- or odd-mode cases. As a consequence, these effective per-unit-length parameters can be explicitly derived in terms of the four elements of the de-embedded finite-cell PNCML section above, i.e. $A^{e,o}$, $B^{e,o}$, $C^{e,o}$ and $D^{e,o}$, in which the subscript e or o corresponds with the even- or odd-mode.

$$Z_{0e,o} = \sqrt{\frac{B^{e,o}}{C^{e,o}}} \quad (5.1)$$

$$\cos(\beta_{e,o}L_p) = \frac{A^{e,o} + D^{e,o}}{2} \quad (5.2)$$

In comparison to the other numerical de-embedding techniques, there exist the two advantageous features of this SOC scheme used here: (i) no pre-requirement in

knowing the even- and odd-mode per-unit-length parameters of the uniform PNCML feed lines in the whole de-embedding procedure; (ii) complete removal of any non-ideal source effects caused by impressed voltages in the source-type MoM.

5.3 UNIFORM CML FOR NUMERICAL VALIDATION

Traditionally, half a cross-section of a symmetrical CML has been taken into account, e.g., [6], in order to realize the impedance matching between either of coupled lines and its corresponding feed line for design of microstrip directional couplers. But, herein, the characteristic impedances of even- and odd-mode are defined through its whole cross-section for more general consideration of wave propagation along these three-conductor transmission lines under even and odd excitations, respectively.

Fig. 5.3(a) and Fig. 5.3(b) depict the cross-section of the whole and half a symmetrical CML transmission line and their relevant definitions of equivalent current and voltage quantities of two dominant modes. For the case of even-mode excitation, the symmetrical plane exactly operates as a perfect magnetic wall (M.W.). In this way, the current at both left- and right-side strip conductors flow in with

Fig. 5.3. The whole and half a symmetrical cross-section of CML structure and their distinctive definitions of equivalent current and voltage quantities for even- and odd-mode. (a) Even-mode: whole. (b) Odd-mode: whole. (c) Even-mode: half. (d) Odd-mode: half.

the same amplitude $+I_e$, thus resulting that the current at the ground plane flows out becomes $-2I_e$. In general, the equivalent even-mode current quantity may be defined as the total current flowing along the two upper strip conductors, i.e. $2I_e$, while its related voltage quantity can be simply expressed by the voltage between the strip conductor and ground, i.e. V_e .

On the contrary, in the case of odd-mode, the odd-symmetrical field distribution may be excited with a perfect electric wall (E. W.) at the central plane, in which the current flowing on the ground plane is canceled by two currents with opposite polarity, i.e. $-I_o$ and $+I_o$. In this case, the odd-mode current quantity is defined as the current at a single strip conductor, i.e. I_o , while the voltage between the two strip conductors, $2V_o$, is the relevant voltage quantity. As compared to the above definitions, the current and voltage quantities of half a CML structure can be simply defined as I_o and V_o , relevant to either of the upper strip conductors, as expressed in Fig. 5.3(c) and Fig. 5.3(d). By looking at Fig. 5.3(a) with Fig. 5.3(c) and Fig. 5.3(b) with Fig. 5.3(d) together, one can figure out the relation between these two sets of even- and odd-mode characteristic impedances, that is,

$$Z_{0e} = Z_{0e}^h / 2 \quad (5.3)$$

$$Z_{0o} = 2Z_{0o}^h \quad (5.4)$$

where the superscript, h, indicates half a CML case. In the following, the former sets of impedance definitions, i.e. Z_{0e} and Z_{0o} , are thoroughly utilized to describe the even- and odd-mode characteristic impedances.

TABLE 5.1
CONVERGENCE BEHAVIOR & COMPARATIVE VERIFICATION OF
DERIVED Z_{0e} (Ω) OF A UNIFORM CML

Freq. (GHz)	MoM-SOC			LineCalc [115]
	$\Delta x=0.3\text{mm}$	$\Delta x=0.2\text{mm}$	$\Delta x=0.1\text{mm}$	
2	29.794	29.375	28.996	28.124
4	29.706	29.287	28.907	28.128
6	29.674	29.255	28.874	28.208
8	29.691	29.271	28.891	28.362
10	29.816	29.395	29.013	28.585
12	29.956	29.535	29.151	28.868
14	30.183	29.761	29.376	29.202
16	30.596	30.174	29.787	29.574
18	31.360	30.937	30.549	29.974
20	32.706	32.282	31.890	30.391

$W=0.6\text{mm}$, $S=0.6\text{mm}$, $\Delta y=0.2\text{mm}$

Table 5.1 shows the convergence behavior of uniform CML Z_{0e} versus different transverse mesh size (Δx) in MoM in the frequency range of 2.0 to 20.0 GHz, in which the longitudinal mesh size (Δy) is fixed as 0.2mm.

As Δx decreases from 0.3 to 0.2 and 0.1mm, the values at different frequencies gradually become small and close to the results from the two-dimensional (2-D) ADS-LineCalc software [115] with the maximum discrepancy of about 2.4%, as can be observed from Table 5.1. Fig. 5.4(a) and Fig. 5.4(b) depict the numerical convergence with aspect to the feed line length (L_f) for the uniform CML characteristic impedance (Z_{0e} & Z_{0o}) and normalized phase constant (β_e/k_0 , β_o/k_0) at the minimum and maximum frequencies of our considered wide range, i.e. $f=2.0$ and 20.0 GHz. As the length (L_f) is extended beyond 3.0 mm, it can be seen that all of quantities seem to converge to their corresponding eventual values. Furthermore, Fig. 5.5(a) and Fig. 5.5(b) are prepared to give a quantitative comparison among our 3-D extracted per-unit-length parameters and those from the

Fig. 5.4. Numerical convergence versus feed line length (L_f) of the uniform CML.

ADS-LineCalc over the frequency range (2.0 to 20.0GHz). Both results are well matched with each other for the even-mode case. However, the odd-mode parameters have the visible discrepancy of about 2.7% for Z_{0o} and 0.5% for β_o/k_0 .

5.4 PNCML: EVEN- AND ODD-MODE

Fig. 5.1(b) to Fig. 5.1(d) describe the three PNCML structures that are periodically loaded by the outer slits with the depth of d_1 , inner slits with the depth

Fig. 5.5. Frequency-dependent per-unit-length transmission parameters of the uniform CML transmission lines as comparison to those derived from the Agilent LineCalc. (a) Characteristic impedance. (b) Phase constant.

of d_2 and bilateral slits of the parallel-coupled strip conductors, namely, PNCML (A), (B) and (C), respectively. T is the periodicity of a single unit cell and t is the width of each slit. Based on the above-described MoM-SOC technique, the per-unit-length parameters of these PNCML transmission lines and their corresponding uniform CML are extracted so as to exhibit the distinctive guided-wave

(a)

(b)

Fig. 5.6. Frequency-dependent per-unit-length transmission parameters of the three PNCML and uniform CML transmission lines: even-mode case. (a) Characteristic impedance. (b) Phase constant.

characteristics of even- and odd-mode propagating along the infinite-extended PNCML.

Fig. 5.6 and Fig. 5.7 depict the frequency-dependent normalized phase constant (β_e/k_0 , β_o/k_0) and characteristic impedance (Z_{0e} , Z_{0o}) of these three PNCML against those of the uniform CML for the even and odd modes, respectively.

Fig. 5.7. Frequency-dependent per-unit-length transmission parameters of the three PNCML and uniform CML transmission lines: odd-mode case. (a) Characteristic impedance. (b) Phase constant.

From Fig. 5.6(a), among the four curves, Z_{0e} of the PNCML (A) with the outer slits is found to achieve the maximum value of about 36.8 Ω at 2 GHz. Meanwhile, Z_{0o} of the PNCML (B) with the inner slits has the largest value of about 91.4 Ω at 2 GHz as seen in Fig. 5.7(a). These results imply that the PNCML (A) and (B) have the weakest coupling extent between the strip-to-ground and strip-to-strip, respectively. Moreover, the actual coupling coefficient between the two coupled

Fig. 5.8. Periodicity-dependent per-unit-length transmission parameters of the three PNCML and uniform CML transmission lines: even-mode case. (a) Characteristic impedance. (b) Phase constant.

microstrip lines can be quantitatively evaluated via $(Z_{0e}^h - Z_{0o}^h)/(Z_{0e}^h + Z_{0o}^h)$ [6], thus illustrating that the PNCML (A) has the tightest coupling.

On the other hand, the even-mode β_e/k_0 is seen from Fig. 5.6(b) to shift up from 2.75, 2.89, 2.90 to 3.05 as the slit loading is changed from disappearance (uniform CML), bilateral, inner to outer sides of coupled strips. It implies that the even-mode

Fig. 5.9. Periodicity-dependent per-unit-length transmission parameters of the three PNCML and uniform CML transmission lines: odd-mode case. (a) Characteristic impedance. (b) Phase constant.

of the PNCML (A) has the largest slow-wave factor. In parallel, Fig. 5.7(b) shows that the PNCML (B) achieves the maximum slow-wave factor, i.e. $\beta_o / k_0 = 2.96$.

From the study here, we figure out that the propagating velocities of the even and odd modes in the CML can be adjusted by periodically loading the inner and/or outer slits. In order to explore the high-directivity directional coupler [46] and harmonic suppressed bandpass filter [47], these two velocities may be equalized by

Fig. 5.10. Comparison between S-parameter magnitudes, $|S_{11}|$ and $|S_{21}|$, of the 3-cell PNCML circuit from the extracted per-unit-length parameters via transmission line theorem and the Momentum simulator. (a) Even-mode case. (b) Odd-mode case.

utilizing the PNCML (B) with inner slits. Looking at Fig. 5.6(b) and Fig. 5.7(b) together, this condition can be easily realized if the slit depth (d_2) is slightly reduced.

Fig. 5.8 and Fig. 5.9 describe the periodicity-dependent per-unit-length parameters of the uniform CML and the three PNCML transmission lines under the fixed slit depths as discussed in Fig. 5.6 and Fig. 5.7. As the periodicity (T) is reduced, all the

parameters tend to rise up consistently in an accelerated manner. Otherwise, they gradually converge to the stable values. Again, the even-mode Z_{0e} of the PNCML (A) is much larger than that of the other two PNCML while the odd-mode Z_{0o} of the PNCML (B) exceeds others. Quick increment of both β_e/k_0 and β_o/k_0 with the reduced T brings out an attractive feature in building up the size-miniaturized CML circuit blocks such as filters and couplers.

In order to make an evident validation of the extracted even- and odd-mode per-unit-length parameters as illustrated in Fig. 5.6 to Fig. 5.9, the PNCML circuit with the three unit cells are constructed and connected with the two uniform CML feed lines at the two sides. Its S-parameters for both even- and odd-mode can be calculated via cascaded transmission line theorem. It should be particularly emphasized in our analysis that only Z_{0e} and Z_{0o} of the uniform CML feed lines, and β_e , β_o , Z_{0e} , Z_{0o} plus the total length $L_p = 3T$ of the finite-cell PNCML under consideration are required. In the Momentum-based simulation, the S-parameters at the two excited ports are simulated and then transferred to those defined at the two terminals of the three-cell PNCML, as shown at the central schematic of Fig. 5.10(b), for quantitative comparison. Fig. 5.10 and Fig. 5.11 illustrate the simulated even- and odd-mode S-parameters of the PNCML circuit with outer slits, i.e. PNCML (A), together with those obtained from the Momentum simulator. They are found in good agreement with each other in both magnitude and phase over the frequency range of 2.0 to 20.0 GHz for both even- and odd-mode cases.

As compared with the CPU time of a few seconds via equivalent transmission line model with the known per-unit-length parameters, the Momentum simulation takes about 20 minutes to derive the results in Fig. 5.10 and Fig. 5.11 with the choice of

(a)

(b)

Fig. 5.11. Comparison between S-parameter phases, Φ_{11} and Φ_{21} , of the 3-cell PNCML circuit from the extracted per-unit-length parameters via transmission line theorem and the Momentum simulator. (a) Even-mode case. (b) Odd-mode case.

30 unit cells per wavelength at the maximum frequency of 20.0 GHz. As a result, this design example not only provides us with comparative validation, but also shows us the usefulness of the extracted per-unit-length parameters in efficiently designing the PNCML-based circuit blocks.

5.5 CONCLUSION

In this chapter, periodically nonuniform coupled microstrip lines (PNCML) are thoroughly characterized as an equivalent uniform coupled transmission line via self-calibrated MoM-SOC technique. For the first time, both two effective per-unit-length transmission parameters, i.e. phase constants and characteristic impedances, are extracted to demonstrate the frequency-dispersive and periodicity-dependent guided-wave characteristics of the two dominant propagating modes. These results may be very useful for accurate design of advanced microstrip couplers and filters with the use of such PNCML structures. SOC-extracted parameters are evidently validated by comparing the transmission-line-simulated S-parameters with those from the Momentum simulator for both even- and odd-mode.

CHAPTER 6

Periodically Nonuniform Coupled Microstrip Line Filters with Harmonic Suppression

6.1 INTRODUCTION

Coupled microstrip line has been widely used as a promising coupling element in the design of bandpass filters due to its several attractive features, such as compact size, low profile and tightened capacitive coupling [13], [104]. However, these microstrip bandpass filters with uniform coupled microstrip line sections usually suffer from the spurious passband at the 2nd resonant frequency of the microstrip line resonator. Consequently, it makes the upper stopband performance worse. As well known today, it is predominantly caused by the non-synchronous feature of even- and odd-mode velocities of propagation in the inhomogeneous dielectric medium. To circumvent this problem, much effort has been made so far to equalize the phase velocities of even- and odd-modes by differentiating their traveling routes [47]–[54], [116]–[118], [122]–[124]. Initially, a simple over-coupled coupled

microstrip line section is formed in [48] to compensate the difference in phase velocity between these two dominant modes. The over-coupling is then introduced to the end stages only [122] and the capacitive termination is realized at each end stages for the harmonic suppression [123]. Alternatively, the nonuniform coupled microstrip line with multiple cascaded sections [116] is made, the meandered parallel coupled line is designed [124] and the dielectric substrate is mechanically suspended above the ground plane [52], [53] to achieve phase compensation.

In recent years, much more interest has been aroused to utilize the periodically nonuniform coupled microstrip line (PNCML), with various unit-cell configurations as a simple and effective approach to address this problematic issue. The sinusoidal modulation on the widths of two strip conductors [49] and the corrugated coupled-slot structure [47], [117] are formulated to extend the actual odd-mode traveling path towards its even-mode counterpart at the 1st harmonic passband. Similarly, the squared grooves are periodically and symmetrically etched out at the both sides of coupled lines [50]. Also the split ring resonators are periodically placed in the proximity of the parallel strip conductors [118]. In [51], the meander coupled microstrip line is constructed with reduced size. In addition, the aperture is periodically formed on the ground plane to equalize the even/odd-mode phase velocities in the design of the wide-band bandpass filters by making use of its tight coupling [54].

As detailed in Chapter 2 and 5, our developed hybrid method of moments (MoM) and short-open calibration (SOC) technique namely, “MoM-SOC”, is extended to directly extract the two per-unit-length transmission parameters (characteristic impedance and phase constant) of PNCML from the full-wave modeling of these

PNCML with finite length. Obtained results of the uniform coupled lines are at first confirmed with the quasi-static ones [125]. These parameters are extracted to expose the basic guided-wave characteristics of the PNCML loaded with transverse slit in a wide frequency range. As such, the exhibited frequency dispersion of these two per-unit-length parameters allows one not only to understand the operating principle of these filters but also to optimize their bandpass filtering behavior via simple and efficient synthesis procedure [45].

6.2 EQUALIZATION OF THE EVEN/ODD-MODE PHASE VELOCITIES

In this section, our effort is primarily made to accurately predict the location of transmission zero for the finitely-extended PNCML section (Type-B in Chapter 5) and further allocate it suitably to suppress the 1st-harmonic passband in the design of microstrip bandpass filters. One- and two-stage PNCML filter design examples are firstly constructed on the RT/duroid 6010LM microwave laminates ($\epsilon_r=10.8$, $h=0.635\text{mm}$). The predicted S -parameters are evidently confirmed by the experimental results.

6.2.1 EXTRACTED PER-UNIT-LENGTH TRANSMISSION PARAMETERS

Based on the above MoM-SOC technique in chapter 5, the finite-extended PNCML line with varied depth (d) of transverse slits in periodical interval is characterized again here for the design of bandpass filter. Fig. 6.1(a) depicts the cross-section of the uniform coupled microstrip line (CML), in which the two parallel-strip conductors are transversely spaced by the gap width (s). Fig. 6.1(b) indicates the geometrical sketch of the finite-extended PNCML loaded with transverse slits in periodical interval. This PNCML is fed at two sides by the two

Fig. 6.1. (a) Cross-section of uniform coupled microstrip line (CML). (b) Top-view of periodically nonuniform coupled microstrip line (PNCML) with finite length.

uniform coupled-line feeders The two per-unit-length transmission parameters are numerically extracted to expose its guided-wave characteristics in comparison to those of the uniform coupled microstrip line (CML).

Fig. 6.2(a) and Fig. 6.2(b) depict the extracted characteristic impedance and normalized phase constant of the finite-length CML with $d=0\text{mm}$ and the PNCML line with $d=0.4\text{mm}$ under the fixed periodicity of $T=1.0\text{mm}$, respectively.

As shown in Fig. 6.2(a), the even- and odd-mode characteristic impedances of this PNCML rise up from 62.4 to 67.9Ω and from 34.5 to 44.2Ω in the non-synchronous way, respectively, as d increases from 0.0mm (uniform) to 0.4mm . Meanwhile, it can be observed in Fig. 6.2(b) that the normalized odd-mode phase constant (β_o/k_0) increases significantly from 2.48 to 3.02 at 2.0 GHz , whereas its even-mode counterpart (β_e/k_0) slightly rises up from 2.82 to 2.98 . In fact, for the uniform CML, a large portion of energy is mainly concentrated in the central gap region for the odd-mode and moves to be distributed in the dielectric layer underneath the two strips for the even-mode [47]. Thus, in PNCML, as the odd-mode traveling path is enlarged, the effective phase velocity of the odd-mode becomes low and its corresponding effective phase constant increases more dramatically while the even-

mode phase constant is almost stably unchanged. In addition, the results with circle markers, in Fig. 6.2, are derived in [125] for the uniform CML case ($d=0.0\text{mm}$) and they are found well matched to those from our MoM-SOC.

Fig. 6.2. Extracted even- and odd-mode per-unit-length transmission parameters of uniform CML and PNCML ($\epsilon_r=10.8$, $h=0.635\text{mm}$, $w=0.6\text{mm}$, $s=0.2\text{mm}$, $T=1.0\text{mm}$, $t=0.2\text{mm}$). (a) Characteristic impedances. (b) Normalized phase constants.

6.2.2 TRANSMISSION ZERO OF PNCML

Following the pioneered work [48], the coupling zero between the two coupled lines with the finite length (L) in the uniform CML or PNCML may be produced at a

certain frequency, i.e. transmission zero. Its location can be explicitly determined in the closed-form equation (6.1), in which the two sets of per-unit-length parameters for even and odd modes are extracted above.

$$\frac{Z_{0e}}{Z_{0o}} = \frac{\sin \beta_e L}{\sin \beta_o L} \quad (6.1)$$

Fig. 6.3 plots the frequency-dependent graphs of the two functions, $Z_{0o}\sin\beta_e L$ and $Z_{0e}\sin\beta_o L$, for adaptive allocation of the transmission zero by adjusting the periodicity (T) and/or slit depth (d). Graphically, the transmission zero can be solved as the intersection point of the two relevant curves. As d increases from 0.0 (uniform CML) to 0.4mm, the transmission zero is found to reduce from 9.92 to 6.95GHz to a great extent. This property will be utilized later on to effectively cancel the 1st-harmonic passband of the CML filter by properly allocating the transmission zero towards the 2nd-resonance of a half-wavelength ML resonator [47]–[51].

Fig. 6.3. Frequency-dependent graphs for adaptive allocation of the transmission zero to the desired frequency. ($\epsilon_r=10.8$, $h=0.635\text{mm}$, $w=0.6\text{mm}$, $s=0.2\text{mm}$, $L=7.0\text{mm}$, $T=1.0\text{mm}$, $t=0.2\text{mm}$).

Fig. 6.4. Normalized phase constants versus ratio t/T . ($\epsilon_r=10.8$, $h=0.635\text{mm}$, $w=0.6\text{mm}$, $s=0.2\text{mm}$, $f=6.95\text{GHz}$).

Fig. 6.4 shows the normalized phase constants as a function of the aspect ratio t/T at the frequency of $f=6.95\text{GHz}$ that is the intersected point of the two solid curves in Fig. 6.3. As the slit with $d=0.4\text{mm}$ is periodically etched out, both phase constants are simultaneously raised due to the well-known slow-wave property. As shown in Fig. 6.4, the odd-mode β_o/k_0 rises up with t/T much more quickly than its even-mode β_e/k_0 , thereby producing the intersection point of these two curves or equalizing these two phase constants around $t/T=0.2$.

6.2.3 FILTER DESIGN EXAMPLES

In this section, the above transmission-zero allocation technique is implemented to explore the harmonic-rejected PNCML bandpass filter. It is realized by suitably allocating the transmission zero to the 1st-harmonic passband as demonstrated in the above sections. Fig. 6.5(a) and 6.5(b) show the layouts of the one-stage CML/PNCML bandpass filters, which are made up of the two cascaded CML/PNCML sections. Relevant S-parameters of these filters are simply calculated via cascaded transmission line theorem, as shown in Fig. 6.6. In the uniform CML

Fig. 6.5. Layouts of the two one-stage bandpass filters. (a) Uniform CML filter. (b) PNCML filter. ($\epsilon_r=10.8$, $h=0.635\text{mm}$, $w=0.6\text{mm}$, $s=0.2\text{mm}$, $L=7.0\text{mm}$, $T=1.0\text{mm}$ and $t=0.2\text{mm}$).

Fig. 6.6. Predicted S-parameters of the one-stage CML and PNCML filters ($\epsilon_r=10.8$, $h=0.635\text{mm}$, $w=0.6\text{mm}$, $s=0.2\text{mm}$, $L=7.0\text{mm}$, $T=1.0\text{mm}$ and $t=0.2\text{mm}$).

filter, the transmission zero of $f_b^0=9.92\text{GHz}$ is higher than its 2nd transmission pole of $f_b^h=8.15\text{GHz}$, thus suffering from the 1st-harmonic passband. But, the 1st-harmonic passband of the PNCML filter is completely canceled since the transmission zero (f_a^0) and 2nd pole (f_a^h) are tuned to the same frequency, i.e. 7.15GHz .

Next, the one- and two-stage bandpass filters are optimally designed to have the same fundamental passband at the frequency of 3.6GHz . Fig. 6.7 illustrates the photograph and frequency responses of the fabricated one-stage filter. The measured

Fig. 6.7. Comparison between the predicted and measured S-parameters of the one-stage PNCML filter. (a) Photograph. (b) S-parameters ($\epsilon_r=10.8$, $h=0.635\text{mm}$, $w=0.6\text{mm}$, $s=0.2\text{mm}$, $T=1.0\text{mm}$ and $t=0.2\text{mm}$, $d=0.38\text{mm}$, $L=7.0\text{mm}$).

results are in good agreement with the network-derived results, thus evidently verifying that the proposed technique can really suppress the 1st-harmonic passband that harmfully existed in the traditional parallel-coupled microstrip line bandpass filter [45].

Finally, a two-stage bandpass filter example is further designed and fabricated. Fig. 6.8(a) is the photograph of this filter and Fig. 6.8(b) shows the comparison between the predicted and measured results over a wide frequency range. Again, both them are well matched with each other. In particular, the measured insertion loss is really raised beyond 45 dB at the 1st harmonic passband around 7.20GHz that is existed in the traditional uniform CML filter.

Fig. 6.8. Comparison between the predicted and measured S-parameters of the two-stage PNCML filter. (a) Photograph. (b) S-parameters. ($\epsilon_r=10.8$, $h=0.635\text{mm}$, 1st-stage PNCML: $w=0.6\text{mm}$, $s=0.2\text{mm}$, $T=1.0\text{mm}$, $t=0.2\text{mm}$, $d=0.375\text{mm}$ and $L=7.0\text{mm}$, 2nd-stage PNCML: $w=1.0\text{mm}$, $s=0.6\text{mm}$, $T=1.0\text{mm}$, $t=0.2\text{mm}$, $d=0.665\text{mm}$ and $L=7.0\text{mm}$).

6.2.4 CONCLUSION

In this work, guide-wave characteristics of the PNCML structure with multiple transverse slits in periodical intervals are thoroughly investigated in terms of two effective per-unit-length transmission parameters. Our effort is primarily made to predict the location of transmission zero of the finitely-extended PNCML section and further allocate it to suppress the 1st-harmonic passband in the design of microstrip bandpass filters. Optimized results with good harmonic suppression are evidently verified over a wide frequency range. Moreover, this PNCML filter has the compact size due to slow-wave propagation of PNCML even and odd-modes.

6. 3 EQUIVALENT J-INVERTER NETWORK FOR PNCML BANDPASS FILTERS WITH HARMONIC SUPPRESSION

Without knowing the even- and odd-mode characteristic impedances and phase velocities of these periodically nonuniform coupled microstrip lines, all the above filters can only be designed optimally. This is done by executing the direct electromagnetic simulation of these electrically large overall layouts at the cost of intensive CPU time and huge memory.

In fact, if the frequency-dispersive coupling performance of these periodically nonuniform coupled microstrip lines is comprehensively investigated, the task of designing these filters may be easily accomplished using the efficiently network-based synthesis approach [13], [104]. In Chapter 5, the effective per-unit-length transmission parameters of infinite-extended periodically nonuniform coupled microstrip lines have been extracted to demonstrate the frequency-dispersive and slow-wave behavior of even and odd dominant modes via MoM-SOC. As an application, the optimized BPF design examples with good harmonic suppression are implemented in Section 6.2. In this section, the two-port periodically nonuniform coupled microstrip line with finite length is characterized as an equivalent J-inverter network, showing its frequency-dispersive coupling performance and reallocating the transmission zero towards suppressing the 1st-harmonic passband of coupled microstrip line filters. On a basis of J-inverter technique, two one-stage bandpass filters are designed, fabricated and measured to quantitatively demonstrate that the 1st-harmonic resonance can be fully suppressed for both tight and weak coupling cases. From a designed example, a three-stage bandpass filter is built and predicted S-parameters are well assured by our experiment over a frequency range.

6.3.1 ONE-STAGE BANDPASS FILTER

Fig. 6.9(a) depicts the geometry of a one-stage bandpass filter, which is constructed by cascading the two identical quarter-wavelength periodically nonuniform coupled microstrip line sections at the central interface (#2). In this aspect, a number of rectangular slits or notches are periodically etched out in the inner edges of the two strip conductors. By adjusting the number and dimension of these slits properly, the even- and odd-mode phase velocities may be equalized at the second resonance of the middle resonator. As shown in Fig. 6.9(a), each length ($L_p=NT$) consists of finite periodical cells (N) with the periodicity (T) and slit width/depth (t & d), whereas W and S are the widths of the two strips and coupling gap. Its equivalent circuit network is described in Fig. 6.9(b). Each section is

Fig. 6.9. (a) Cross-section of uniform coupled microstrip line (CML). (b) Top-view of periodically nonuniform coupled microstrip line (PNCML) with finite length.

modeled as a J-inverter network with the susceptance (J) at center and two equal electrical lengths ($\theta/2$) at the two sides. As studied in the uniform coupled microstrip line case [119], the distributed J-susceptance has been shown to vary as a quasi-periodical function of frequency. It reaches the peak and null values at the frequencies of $\theta/2=90^\circ$ and $\theta/2=180^\circ$ respectively.

6.3.2 EQUIVALENT J-INVERTER NETWORK

As detailed in Chapter 5, the characteristic impedances (Z_{0e} , Z_{0o}) and phase constants (β_e , β_o) of the periodically nonuniform coupled microstrip line can be first effectively extracted using our developed full-wave MoM-SOC technique. As a result, this allows one to derive the two-port impedance (Z-) matrix of the open-circuited periodically nonuniform coupled microstrip line section with the length of $L_p = NT$ in terms of even and odd-mode per-unit-length parameters of a generalized coupled transmission line in an inhomogeneous medium in such a closed form [120].

$$Z_{11} = Z_{22} = \frac{-j}{2}(Z_{0e}^h \cot \beta_e L_p + Z_{0o}^h \cot \beta_o L_p) \quad (6.2)$$

$$Z_{12} = Z_{21} = \frac{-j}{2}(Z_{0e}^h \csc \beta_e L_p - Z_{0o}^h \csc \beta_o L_p) \quad (6.3)$$

The superscript, h, indicates half a symmetrical coupled microstrip line case in conjunction with the definition in [13], [104].

Due to the different definitions of wave current and voltage quantities in the modeling of the whole and half a symmetrical coupled microstrip line, the above two sets of even- and odd-mode characteristic impedances can be explicitly related as,

$$Z_{0e}^h = 2Z_{0e} \quad (6.4)$$

$$Z_{0o}^h = Z_{0o} / 2 \quad (6.5)$$

Equations (6.2) and (6.3) imply that each element in the Z-matrix is purely imaginary and they can be converted into the relevant admittance (Y-) matrix with

the two independent susceptances of $B_{11}=B_{22}$ and $B_{12}=B_{21}$. Under the equivalence of two matrices or networks for the same periodically nonuniform coupled microstrip line block, the J-susceptance (J) and electrical length (θ) can be expressed as [119],

$$\bar{J} = \frac{\tan(\theta/2) - \bar{B}_{11}}{\bar{B}_{12} \tan(\theta/2)} \quad (6.6)$$

$$\theta = n\pi + \tan^{-1} \left\{ \frac{2\bar{B}_{11}}{1 - \bar{B}_{11}^2 + \bar{B}_{12}^2} \right\} \quad (6.7)$$

where $\bar{J} = J/Y_0$, $\bar{B}_{11} = B_{11}/Y_0$, $\bar{B}_{12} = B_{12}/Y_0$, n is an integer number and Y_0 is the characteristic admittance of the uniform lines that excite the open-circuited periodically nonuniform coupled microstrip line at the two sides.

6.3.3 FREQUENCY-DISPERSIVE J-INVERTER NETWORK PARAMETERS

As the infinite-long periodically nonuniform coupled microstrip line is modeled via MoM-SOC, done in Chapter 5, the two J-inverter network parameters can be calculated using Equations (6.2) to (6.7). Fig. 6.10(a) and 6.10(b) describe the three sets of normalized J-inverter susceptance ($\bar{J}$) and electrical line length ($\theta/2$) versus frequency under three different depths of rectangular slits, i.e. $d=0.00$ (uniform case), 0.36, 0.50 mm. In comparison with the approximate assumption of no frequency dispersion, the dispersive coupling performance of this periodically nonuniform coupled microstrip line is characterized as a simple equivalent J-inverter network. As seen in Fig. 6.10(a), the parameter $\bar{J}$ increases quickly, reaches the peak, falls down to the null and then goes up again. It exhibits a non-monotonous and quasi-periodical coupling behavior as a function of frequency. The coupling degree at the frequency, f_a^0 , f_b^0 or f_c^0 , becomes null, thus it brings out the

Fig. 6.10. Comparison between the predicted and measured S-parameters of the one-stage PNCML filter. (a) Photograph. (b) S-parameters ($\epsilon_r=10.8$, $h=0.635\text{mm}$, $w=0.6\text{mm}$, $s=0.2\text{mm}$, $T=1.0\text{mm}$ and $t=0.2\text{mm}$, $d=0.38\text{mm}$, $L=7.0\text{mm}$).

transmission zero in the S-parameters.

The electrical line length ($\theta/2$) at the two sides of the J-susceptance is found to quasi-linearly go up from 33° to 273° as the frequency increases from 1.0 to 7.0 GHz. From Fig. 6.9, it can be figured out that the middle transmission line resonator is actually formed by cascading two identical electrical lengths ($\theta/2$). Its overall length is equal to θ . Intuitively, the transmission line resonator resonates at the

frequencies of $\theta=180^0$, 360^0 , 540^0 and so on. The first resonance happened at $\theta=180^0$ is usually utilized to make up the dominant passband in the design of a bandpass filter. The second resonance at $\theta=360^0$ ($\theta/2=180^0$ in Fig. 6.10(b)) contributes to the 1st harmonic spurious passband with the central frequencies, i.e. f_a^h , f_b^h or f_c^h , in conjunction with the three different slit depths (d) as shown in Fig 6.10(b). In fact, these three depths (d) are simultaneously expressed in Fig. 6.10 to comparatively demonstrate the three distinctive cases about the relationships between the transmission zero and 1st-harmonic resonance frequencies, that is,

$$\begin{cases} f_a^0 < f_a^h, & d = 0.50 \text{ mm} \\ f_b^0 = f_b^h, & d = 0.36 \text{ mm} \\ f_c^0 > f_c^h, & d = 0.00 \text{ mm} \end{cases} \quad (6.8)$$

where the superscripts 0 and h indicate the transmission zero and 1st-harmonic resonance cases, respectively.

6.3.4 TRANSMISSION ZERO REALLOCATION

To consider the tight and weak coupling cases as requested in the multiple-stage bandpass filter, both the transmission zeros and harmonic resonant frequencies, also the frequency quantities, f^0 and f^h , are plotted in Fig. 6.11 with respect to the slit depth (d) for S=0.2 mm and 0.6 mm, respectively. By adjusting the slit depth (d) under the pre-selected periodicity (T), these two curves or lines are intercepted with each other at a certain frequency where $f^0 = f^h$.

Furthermore, as shown in Fig. 6.11, the intersection point occurs at d=0.36 mm for the tight coupling case with S=0.2 mm and T=1.2 mm, whereas it appears at d=0.39 mm for the weak coupling case with S=0.6 mm and T=0.68 mm.

Fig. 6.11. Frequency of transmission zero and harmonic resonance versus the depth of transverse slit. (a) Tight coupling with $S=0.2$ mm. (b) Weak coupling with $S=0.6$ mm.

Of course, this interception point may also be generated and then reallocated to the particular frequency by reducing the periodicity (T) or increasing the unit number (N) for a fixed coupled length (L_p) properly. Table 6.1 tabulates the two frequency quantities (f^0 and f^h) with respect to unit number (N). The former one largely exceeds the latter one and decreases in a faster manner to meet up with the latter one at $N=9$, where $f^0 = f^h = 4.9$ GHz. As demonstrated later, these two

TABLE 6.1
FREQUENCIES OF TRANSMISSION ZERO AND HARMONIC RESONANCE VERSUS THE NUMBER
OF FINITE CELLS UNDER THE FIXED COUPLED LENGTH (L_p)

N	0	1	2	3	4	5	6
f^0	6.1	5.96	5.81	5.66	5.52	5.38	5.25
f^h	5.45	5.37	5.3	5.23	5.16	5.10	5.05
N	7	8	9	10	11	12	13
f^0	5.12	5.01	4.9	4.78	4.68	4.58	4.48
f^h	4.99	4.94	4.9	4.84	4.80	4.78	4.75

Frequency unit: GHz; substrate: $\epsilon_r=10.8$, $h=1.27$ mm; $W=0.6$ mm, $S=0.2$ mm, $t=0.2$ mm, $L_p=10.8$ mm

equalized frequencies provide us with a direct and effective approach in canceling or suppressing the 1st-harmonic spurious passband in filter design.

6.3.5 HARMONIC-RESONANCE CANCELLATION

Let's study the frequency response of the one-stage bandpass filter in Fig. 6.9(a), as demonstrate whether the 1st-harmonic resonance can be cancelled via the transmission zero reallocation technique presented above. Fig. 6.12 depicts the frequency responses of transmission coefficient ($|S_{21}|$) for the one-stage bandpass filter with the three different slit depths listed in Equation (6.8). As the transmission zero is exactly reallocated to its relevant resonant frequency, i.e. $f^0 = f^h$ ($d = 0.36$ mm), the transmission pole can be cancelled to a great extent. Otherwise, the transmission pole and zero appear separately at the two unequal frequencies in the other two cases in conjunction with $d = 0$ and 0.50 mm. As a result, the appearance of these transmission poles will damage the stopband performance as can be observed in Fig. 6.12.

Fig. 6.12. Frequency responses of one-stage bandpass filters with different slit depths under the fixed coupled length ($L_p=10.8$ mm) and periodicity ($T=1.2$ mm).

Fig. 6.13. Frequency responses of one-stage bandpass filters with different numbers of finite cells under the fixed coupled length ($L_p=10.8$ mm).

Furthermore, the frequency response of this one-stage bandpass filter is investigated versus different slit number (N) under the fixed depth (d) and physical length (L_p). The simulated results are illustrated in Fig. 6.13. On the basis of the derived parameters in Table 6.1, the finite cell number (N) is selected as 0, 5, 9 and 13. This will take into account all the three distinctive cases as summarized in Equation (6.8) as well as the uniform case with $N=0$. At $N=0$, the frequency at the

minimum $|S_{21}|$ appears around 6.1 GHz. It is brought out by the transmission zero (f^0) as given in Table 6.1. It is much higher than that at the maximum $|S_{21}|$, i.e. 1st-harmonic resonant frequency ($f^h = 5.45$ GHz). As N enlarge to 5, the two frequencies of transmission zero and pole is shifted downward in different extents. Thereby reducing their discrepancy ($f^0 - f^h$) from 0.65 GHz to 0.28 GHz. At $N=9$, the zero is reallocated to its pole, i.e. $f^0 - f^h = 0$ GHz. So, the concerned pole or 1st-harmonic resonance is completely suppressed as illustrated in the solid line in Fig. 6.13. When N further increases to 13, the pole emerges again at the frequency ($f^h = 4.75$ GHz) higher than its zero at $f^0 = 4.48$ GHz. It can be observed in Fig. 6.13.

6.3.6 FILTER DESIGN: PREDICTED AND MEASURED RESULTS

On the basis of the above discussion, the two extreme cases with tight and weak coupling are considered initially to make up the two one-stage bandpass filters with harmonic suppression at their 1st-order spurious passbands. Fig. 6.14(a) and 6.14(b) are the photographs of the two fabricated filters with narrow and wide coupling gap widths (S) respectively. Our design goal is to achieve identical dominant resonant frequencies at 2.45 GHz while canceling the 1st-harmonic resonance in both cases.

Fig. 6.14. Photographs of the two fabricated one-stage bandpass filters. (a) Tight coupling. (b) Weak coupling.

Fig. 6.15. Predicted and measured S-parameters of the one-stage bandpass filters that are shown in Fig. 7.6. (a) Tight coupling with $S=0.2 \text{ mm}$. (b) Weak coupling with $S=0.6 \text{ mm}$.

We need to find the proper values for the periodicities and slit depths according to the results in Fig. 6.11 and Table 6.1.

Measured results are plotted together with the predicted results in Fig. 6.15 giving an evident confirmation that the 1st-harmonic resonances are completely suppressed in both tight and weak coupling cases.

With the innovative use of these periodically nonuniform coupled microstrip lines, a multi-stage bandpass filter with 1st-harmonic suppression will be able to be

Fig. 6.16. Photograph of the fabricated three-stage bandpass filter.

Fig. 6.17. Predicted and measured S-parameters of the three-stage bandpass filter. (tight-coupling section: $W=0.6$ mm, $S=0.2$ mm, $T=1.2$ mm, $d=0.38$ mm and $N=9$; weak-coupling section: $W=1.18$ mm, $S=0.5$ mm, $T=0.6$ mm, $d=0.45$ mm and $N=17$).

constructed in reality. In a design example, a three-stage bandpass filter is designed optimally. The photograph of the fabricated sample is shown in Fig. 6.16.

The tightly coupled section is arranged at the input and output. The weakly coupled one is put at the middle sections. In comparison with the symmetrical geometry of the second resonator, the first and third resonators are constructed by cascading the two dissimilar periodically nonuniform microstrip line sections. The predicted and measured S-parameters are plotted in Fig. 6.17, showing a good agreement with each other in the frequency region from 1.0 to 7.0 GHz. The measured results illustrate that the return loss achieves about 20 dB in the whole

dominant passband at around 2.45 GHz. The insertion loss is higher than 40 dB at the 1st harmonic passband located at around 4.90 GHz.

6.3.7 CONCLUSION

This work has studied the periodically nonuniform coupled microstrip lines with finite length in terms of a generalized coupling-oriented J-inverter network. It does exhibit the basic principle of suppressing the 1st-harmonic passband in filter design with the use of the transmission zero reallocation technique. Extensive results have demonstrated that the transmission zero or J-susceptance null of the periodically nonuniform coupled microstrip line can be reallocated to cancel the 1st-harmonic resonance. This is done by adjusting the periodicity and/or loaded slit dimension properly. The two one-stage bandpass filters are designed and fabricated to provide an initial support on the proposed technique for the periodically nonuniform coupled microstrip line with tight and weak coupling. Lastly, a three-stage bandpass filter is designed optimally and its predicted results are evidently assured from our experiment.

CHAPTER 7

Coupling Dispersion of Parallel-Coupled Microstrip Lines for Dual-Band Microstrip Bandpass Filters

7.1 INTRODUCTION

Multi-band devices, such as multi-band antenna, multi-band filter and multi-band low-noise amplifier, have been recently receiving a tremendously increasing application in exploring many advanced wireless systems with simultaneous operations at multiple frequency bands [55]. As pointed out [56], multi-band passive circuits basically determine the multi-band operation quality, overall size and fabrication cost of a RF and wireless module in the integration technologies. Of them, bandpass filters with multiple passbands are considered as one of the most key components in these multi-band systems. Due to the shortage of matured design procedure, it becomes the most challenging issue for anyone to design the multi-band filters with good passband performances even for a dual-band case.

In [57], the two filtering circuits with different passbands are connected together to implement the initial type of a dual-band filter. But, this solution unfortunately increases the insertion loss and the overall size of a resultant filter block. In [126], [127], the transmission zeros are introduced at the middle passband of wide bandpass filters to enforce the existence of two separate passbands. Recently, the stepped-impedance resonator (SIR) is utilized to make up a filter with dual passbands [59], [60], where the central frequencies are determined by the aspect ratio of the two characteristic impedances. Further, to strengthen the Q factors or lower the insertion loss in both passbands, the two dual-band impedance transformers have to be additionally constructed at the two ports of the core cascaded-resonator section as implemented in [60], [61]. Thus, these dual-band distributed transformers not only significantly enlarge the overall size of the resultant filter block and also bring out the complexity in simultaneously achieving the two operating frequencies, reducing the insertion losses and adjusting their fractional bandwidths (FBWs) in both dual passbands.

In this chapter, the parallel-coupled microstrip lines (PCMLs) with distributed capacitive coupling [119] are characterized and used to inter-link the microstrip SIRs [128] towards build up a miniaturized dual-band filter without needing external dual-band impedance transformers. In this context, each SIR is at first designed to produce the two preferred resonant frequencies and then coupling dispersion of these PCMLs is modeled to achieve the adjustable coupling degrees at these two frequencies by varying the overlapped coupling lengths at each section. After the theoretical investigation of the dual-band PCML and single dual-band SIR circuits, several two-stage dual-band filters are designed to demonstrate the

controllable dual-band widths and predicted results are well conformed by our experiment.

Fig. 7.1. Geometrical diagrams of the stepped-impedance resonator (SIR) and its constituted dual-band bandpass filter (BPF). (a) SIR ($R_Z < 1$). (b) One-stage SIR BPF (coupling length: L_{C1}).

7.2 CHARACTERIZATION OF DUAL-BAND SIR AND PCML

Fig. 7.1(a) depicts the geometry of a microstrip SIR. Its 1st two resonant frequencies, f_0 and f_{s1} , can be properly separated by varying the aspect ratio of the two impedances, i.e. $R_Z = Z_2/Z_1$, under the pre-selected central and the two side sections in this SIR. Following the analysis given in [128], the two resonant frequencies can be solved as a function of impedance ratio (R_Z) and length ratio (L_1/L_2) in terms of Equations (7.1) and (7.2). They are readily allocated to 2.4 and 5.2 GHz, respectively, in this work. Fig. 7.1(b) shows that such a SIR is launched at two sides by the two PCML sections with the overlapped length of L_{C1} . In order to facilitate the design of these PCMLs with long length, the two side-section lengths are stretched as long as possible.

$$R_Z - \tan \theta_1 \tan \theta_2 = 0 \quad (7.1)$$

$$R_Z \tan \theta_1 + \tan \theta_2 = 0 \quad (7.2)$$

7.2.1 COUPLING DISPERSION OF PCML

Fig. 7.2(a) depicts the geometry of the PCML with overlapped coupling length of L_{C1} . Its equivalent circuit may be expressed as a J-inverter susceptance (J) and the two equal electrical lengths ($\theta/2$) with characteristic admittance Y_0 , as shown in Fig. 7.2(b).

Fig. 7.2. Parallel-coupled microstrip line (PCML) to be characterized. (a) Geometry. (b) Equivalent J-inverter network.

As the characteristic impedances (Z_{0e} , Z_{0o}) and phase constants (β_e , β_o) of the even and odd dominant modes in the uniform PCML are characterized, the two-port impedance matrix can be deduced with the four elements as,

$$Z_{11} = Z_{22} = \frac{-j}{2}(Z_{0e} \cot \beta_e L_{C1} + Z_{0o} \cot \beta_o L_{C1}) \quad (7.3)$$

$$Z_{12} = Z_{21} = \frac{-j}{2}(Z_{0e} \csc \beta_e L_{C1} - Z_{0o} \csc \beta_o L_{C1}) \quad (7.4)$$

Under the network equivalence, the two J-inverter network parameters can be obtained in terms of the two independent normalized susceptances, such that

$$\bar{J} = \frac{J}{Y_0} = \frac{\tan(\theta/2) - \bar{B}_{11}}{\bar{B}_{12} \tan(\theta/2)} \quad (7.5)$$

$$\theta = n\pi + \tan^{-1} \left\{ \frac{2\bar{B}_{11}}{1 - \bar{B}_{11}^2 + \bar{B}_{12}^2} \right\} \quad (7.6)$$

where $\bar{B}_{11} = B_{11}/Y_0$, $\bar{B}_{12} = B_{12}/Y_0$, n is an integer number.

Fig. 7.3(a) and Fig. 7.3(b) illustrate the derived normalized J-inverter susceptance ($\bar{J}$) and equivalent electrical length ($\theta/2$) of our interest under different lengths (L_{C1}). The parameter $\bar{J}$ is seen to increase, reach to the maxima and then decrease as a non-monotonic function of the frequency over 1.0 to 7.0 GHz for all the three cases. Further, as the length (L_{C1}) is stretched, the coupling peak is gradually reduced to simultaneously enlarge and reduce the coupling degrees at the two operating frequencies, i.e. $f_0=2.4\text{GHz}$ and $f_{s1}=5.2\text{GHz}$. As L_{C1} is shortened from 9.00, 7.30 to 5.60mm, $\bar{J}$ at 2.4GHz decreases from 0.30, 0.24 to 0.18 while $\bar{J}$ at 5.2 GHz increases from 0.27, 0.35 to 0.36. It implies that the coupling degree of this PCML becomes weaker at the first operating band and stronger at the second

(a)

Fig. 7.3. Extracted J-inverter network parameters of two-port PCML.

operating band. In other words, the fractional bandwidths of both bands may be freely tuned by adjusting the coupling length L_{C1} . Fig. 7.3(b) shows the quasi-linearly increased electrical lengths related to the two resonant frequencies of the dual-band SIR.

7.2.2 PCML-EXCITED DUAL-BAND SIR

Using the above-modeled PCML section, a dual-band SIR circuit is constructed as

Fig. 7.4. S_{21} -magnitude of a PCML-excited dual-band SIR with varied coupling length L_{C1} .

shown in Fig. 7.1(b). Its S_{21} -magnitude under the three different coupling lengths is simulated by the Momentum simulator [115] and plotted in Fig. 7.4 to initially demonstrate the controllable dual-band pass bandwidths at 2.4 and 5.2 GHz. In this aspect, the fractional bandwidth (FBW) observes to be widened from 6.8%, 8.2% to 12% at 2.4GHz-band and narrowed from 10.5%, 8.2% to 5.2% at 5.2GHz-band, respectively. Of importance, these results illustrate that the bandwidths of the dual passbands can be freely adjusted to well realize the three distinctive dual-band cases, as listed in Equation (7.7), which have not yet been reported so far.

$$\begin{cases} FBW_{f_0} > FBW_{f_{s1}}, & L_{C1} = 9.00mm \\ FBW_{f_0} = FBW_{f_{s1}}, & L_{C1} = 7.30mm \\ FBW_{f_0} < FBW_{f_{s1}}, & L_{C1} = 5.60mm \end{cases} \quad (7.7)$$

7.2.3 DUAL-BAND FILTER WITH CONTROLLABLE PASS FBWS

Fig. 7.5 indicates the layout of the proposed two-stage dual-band filter. Its dual-band performances, insertion losses and bandwidths, suppose to be effectively controlled with freedom by adjusting the two coupling lengths, L_{C1} and L_{C2} ,

Fig. 7.5. Layout of the proposed two-stage SIR dual-band filter with controllable fractional pass bandwidths.

between the feed line and SIR as well as two adjacent SIRs. Following the above-discussed three cases, the three dual-band filters are optimally designed to realize the distinctive dual-band performances, i.e. (a) wide FBW at 2.4GHz and narrow

Fig. 7.6. Photograph of the three dual-band SIR filters, type-A, type-B and type-C, with varied dual-passband fractional bandwidths.

FBW at 5.2GHz; (b) equalized FBWs at 2.4 and 5.2GHz; and (c) narrow FBW at 2.4GHz and wide FBW at 5.2GHz.

After optimization design is carried out to minimize the return losses in both passbands, the three dual-band filters are fabricated and their relevant photographs are shown in Fig. 7.6, respectively. As can be seen here, the coupling lengths, L_{C1} and L_{C2} , are selected relatively long in the first filter to make up the PCML with tighter coupling at 2.4GHz and weaker at 5.2GHz as instructed in Fig. 7.2(a). In the second filter, L_{C1} and L_{C2} are properly shortened, aiming at allocate closely the two passband widths. At last, these two lengths are further reduced so as to realize the narrow FBW at 2.4GHz and wide FBW at 5.2GHz due to their weak and strong coupling degrees of the two PCML sections, respectively.

Table 7.1 tabulates the measured dual FBWs at the low and high passbands, FBW_{f0} and FBW_{fs1} , of these three filter samples. One can clearly see that FBW_{f0} is much wider than FBW_{fs1} in the type-A, gradually approaches and then exceeds the latter one in the type-C. It is for the first time confirmed in experiment that in

TABLE 7.1.
MEASURED DUAL-BAND FBWS OF THREE DESIGNED FILTERS

Filter	type-A	type-B	type-C
FBW_{f0}	9.3%	6.5%	4.7%
FBW_{fs1}	3.5%	5.6%	5.6%

addition to the two tunable dual passbands as realized using the SIR [60], the coupling dispersion of the PCML can be further utilized to tune and control the FBWs of the dual operating passbands. Fig. 7.7 plots the predicted and measured S-parameters. They are well matched with each other over a wide range, including the

(a)

(b)

Fig. 7.7. Predicted and measured results of the three dual-band SIR bandpass filters with varied FBWs.

dual passbands.

7.2.4 CONCLUSION

In this work, a new dual-band filter with controllable fractional bandwidths is proposed and constructed by using the distributed PCML coupled-line and dual-resonance SIR. After coupling dispersion of the PCML with varied lengths is studied in terms of explicit J-inverter susceptance, the three SIR circuits are modeled to exhibit the two tunable passbands. Further, the three SIR dual-band filters with two SIRs are optimally designed with varied dual-band FBWs as well verified by our experiment. This simple but effective design procedure provides us a powerful capacity in exploring the dual-band filter without needing any external dual-band matching network [60], [61], thus miniaturizing the overall size and reducing the design complexity.

7.3 COMPACT DUAL-BAND FILTERS

7.3.1 INTRODUCTION

Parallel-coupled microstrip dual-band bandpass filter without any external feeds is originally presented in Section 7.2 to realize the controllable dual-passband bandwidths at the two specified frequencies. To further minimize the overall size as usually requested, a novel compact dual-band microstrip bandpass filter is constructed and implemented in this chapter. In this aspect, a modified half-wavelength SIR resonator [128] with sinuous configuration is presented and designed with the dual passbands operating at 2.4 and 5.2 GHz. The parallel-coupled microstrip line is then characterized in terms of J-inverter network, as done in Chapter 6, to simultaneously allocate its first two coupling peaks at the above two resonant frequencies. As such, good transmission performances are achieved in both dual passbands. Finally, a two-stage dual-band filter is optimally designed and fabricated. Predicted dual-band frequency response is well confirmed by experiment of a fabricated filter block.

7.3.2 PROPOSED COMPACT DUAL-BAND MICROSTRIP FILTER

Fig. 7.8 shows the geometrical schematic of the proposed dual-band microstrip bandpass filter. First of all, the sinuously-shaped SIR resonator is constructed by cascading a short-length high-impedance section in the center with the two long-length low-impedance sections in the two sides. In this design, the widths of those two distinctive sections, i.e. W_1 and W_2 , are pre-selected at 0.2 and 0.4mm. Following the analysis in [128], the resonant conditions of the normal SIR resonator, as shown in Fig. 7.9, can be explicitly established. As a result, the two relevant

Fig. 7.8. Schematic of the compact dual-band microstrip bandpass filter. Substrate: $\epsilon_r=10.8$, thickness=0.635 mm. $L_1=7.6$, $L_2=1.4$, $L_3=3.05$, $L_4=1.8$, $L_5=1.2$, $L_6=8.1$, $L_7=1.55$, $W_1=0.2$, $W_2=0.4$, $S_1=0.2$, $S_2=0.64$ and $S_3=0.9$. All are in mm.

Fig. 7.9. Geometrical diagram of the stepped-impedance resonator (SIR).

resonant frequencies, f_0 and f_{s1} , can be determined as the two solutions of the following two transcendental equations, respectively.

$$R_z - \tan \theta_1 \tan \theta_2 = 0 \quad (\text{at } f=f_0) \quad (7.8)$$

and

$$R_z \tan \theta_1 + \tan \theta_2 = 0 \quad (\text{at } f=f_{s1}) \quad (7.9)$$

where R_z is the ratio of the two characteristic impedances, i.e. $R_z=Z_2/Z_1$. Fig. 7.10 plots the ratio of the first two resonant frequencies, f_{s1}/f_0 , as a function of the impedance ratio (R_z) under three different ratios of the two line lengths in a SIR resonator, i.e. $\theta_2=\theta_1$, θ_2 , and θ_3 . Considering that $f_0=2.4$ and $f_{s1}=5.2$ GHz are

Fig. 7.10. Frequency ratio (f_{s1}/f_0) with respect to impedance ratio ($R_z = Z_2/Z_1$) for a SIR resonator.

considered in this work, R_z can be explicitly determined from Fig. 7.10 if θ_2/θ_1 is specified.

Next, the parallel-coupled microstrip lines are introduced between the feed line and its relevant SIR as well as between the two adjacent SIRs, as illustrated in Fig. 7.8. As discussed in Chapter 6, the coupling degree of this parallel-coupled line is exhibited to vary as a non-monotonic function of frequency. As such, the coupled spacing and lengths can be properly tuned to simultaneously satisfy the coupling degrees and minimize the return losses in both dual passbands without needing any external impedance transformers as inquired in the previous dual-band filters [60], [61]. In this way, the coupling dispersion of parallel-coupled microstrip line is characterized as done in Chapter 6 and then optimally characterized to allocate the first two coupling peaks towards the centers of dual passbands. To do it, the overlapped lengths (L_1 , L_2) and coupled spacing (S_1) need to be characterized to determine the coupling degree between the feed line and the SIR at the input and output. In the meantime, the coupled lengths (L_6 and L_7) and spacing (S_2 , S_3) are

adjusted to control the coupling behavior between the two sinuous SIR resonators. In the final stage, the proposed dual-band filter with the 50- Ω feed lines at two ports is optimally designed with the use of the Agilent Momentum simulator. The target here is to simultaneously achieve the satisfactory transmission performance in the dual passbands by suitably adjusting the above dimensions.

7.3.3 RESULTS AND DISCUSSION

Fig. 7.11 and Fig. 7.12 show the frequency responses of the reflection coefficient $|S_{11}|$ under the different selected coupled spacing (S_2 and S_3) between the two sinuously-shaped SIRs. Herein, all the other dimensions of this filter are unchanged and they are given in Fig. 7.8. From Fig. 7.11, we can see that $|S_{11}|$ in the 2.4-GHz band keeps almost unchanged below -20 dB with two attenuation zeros regardless of varied S_2 . Meanwhile, $|S_{11}|$ in the 5.2-GHz band is immensely varied in shape. As S_2

Fig. 7.11. Frequency responses of $|S_{11}|$ under different coupling spacing S_2 between two sinuous SIR resonators.

is widened from 0.54, 0.64 to 0.74 mm, $|S_{11}|$ in the 2nd-band is suddenly changed in shape from a single attenuation zero to the double ones and the relevant maximum

Fig. 7.12. Frequency responses of $|S_{11}|$ under different coupling spacing S_3 between two sinuous SIR resonators.

value at center is then moved up because of deficient coupling degree. On the other hand, we can find from Fig. 7.12 that the variation in S_3 from 0.8, 0.9 to 1.0 mm coincidentally affects the bandpass performances in both dual bands. It implies that S_2 only dominates the coupling degree in the 2nd-band while S_3 controls the coupling degrees in both bands.

Following the above discussion, our optimization design is carried out by taking into account strip thickness of $17\mu\text{m}$. The design procedure is to at first determine S_3 with satisfactory performance at the 1st-band and then to adjust S_2 towards good transmission in the 2nd-band. The designed filter circuit occupies the overall size of about $10.2 \times 12.3 \text{ mm}^2$. The photograph of the fabricated filter is shown in Fig. 7.13 and its two-port frequency response is then measured using the universal substrate test fixture (WK-3000: supplied by the Inter-Continental Microwave, as shown in Fig. 4.11).

Fig. 7.14 plots the predicted and measured S-parameters in the frequency range covering the concerned dual bands. Both them are found in very reasonable

Fig. 7.13. Photograph of a fabricated dual-band microstrip bandpass filter.

Fig. 7.14. Simulated and measured frequency responses of the optimized dual-band microstrip bandpass filter.

agreement with each other. In the dual passbands, the fractional frequency bandwidths are about 7.0% and 4.0%. The measured return losses in both 2.4GHz- and 5.2GHz-bands are higher than 20 dB while the measured insertion losses are 1.8 dB and 2.9 dB, respectively.

7.3.4 CONCLUSION

A novel compact dual-band microstrip bandpass filter is proposed and constructed by using the sinuously-shaped dual-resonance SIR resonator and dual-band parallel-

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coupled microstrip lines. After a brief description is given on its distinct geometry and optimization design approach, this filter is designed, implemented and fabricated. Without needing any external dual-band impedance-transformer feeds, this proposed dual-band filter has not only significantly miniaturized the overall size, but also achieved the good dual-passband filtering performances at the two specified bands, i.e. 2.4 and 5.2 GHz. The predicted results are finally verified by our experiment of a fabricated dual-band filter.

CHAPTER 8

Novel Dual-Band Filters with Controllable FBWs and Good In-Between Isolation

8.1 INTRODUCTION

Dual-band bandpass filters have been receiving a great interest in the design of advanced wireless communication systems [55], [56], [59]–[62], [126], [127], [129]. To improve the stopband attenuation, coupled resonator pairs are proposed in [62] to generate two transmission zeros in the higher side of each passband. Unfortunately, the filtering performances in the dual passbands are found not so good because of the unexpected parasitic effects and approximate modeling.

To minimize the overall size of circuit and improve the filtering performance in both dual passbands, parallel-coupled microstrip dual-band bandpass filter, as shown in Fig. 8.1(a), is originally presented in Chapter 7 without installing any external dual-band matching networks. The three sets of S_{21} -magnitude are plotted

Fig. 8.1. Geometrical diagrams of the proposed dual-band parallel-coupled microstrip bandpass filters. (a) One-stage filter. (b) Two-stage filter.

in Fig. 7.2 to demonstrate that the dual bandwidths at 2.4 and 5.2 GHz are controllable with respect to the changed coupling lengths (L_{C1}).

In this Chapter, the parallel and anti-parallel coupled microstrip lines [130] are thoroughly investigated and they are utilized to make up a novel microstrip dual-band filter with miniaturized size, controllable dual-passband widths and good in-between isolation. Fig. 8.1(b) shows the geometry of the proposed two-stage dual-band filter. At first, these two coupled microstrip lines are characterized using the self-calibrated full-wave method of moments. Their unified equivalent J-inverter networks are extracted to show their frequency-distributed coupling behaviors [119]. Several two-stage dual-band filters are then designed to exhibit the controllable dual-band performances as demonstrated in both simple network-cascaded prediction and full-wave Momentum simulation. At last, a three-stage dual-band bandpass filter is optimally designed with the improved in-between isolation. The predicted results are further confirmed by experiment.

8.2 COUPLING PROPERTIES OF COUPLED MICROSTRIP LINES

Fig. 8.2(a) and 8.2(b) show the physical configurations of the two parallel-coupled microstrip lines, namely, Type-I and Type-II, to be considered in the modeling platform. These two coupling structures are characterized as the commonly unified equivalent J-inverter network, as shown in Fig. 8.2(c), including a susceptance (J)

Fig. 8.2. Geometrical diagrams and unified equivalent network of the two parallel-coupled microstrip lines to be characterized. (a) Anti-parallel coupled line: Type-I. (b) Pro-parallel coupled line: Type-II. (c) J-inverter network.

at center and two equal electrical line lengths ($\theta/2$) at two sides. These so-called anti-parallel and parallel coupled lines are analyzed in [130] using the simple even- and odd-mode approach towards suppressing the harmonic passbands in the filter

design. To more accurately investigate their coupling behaviors including the parasitic open-end effects, these two-port coupled lines are modeled here using the full-wave admittance-type method of moments (MoM) [86]. Then, the MoM-simulated Y-matrices at the two external ports with the impressed voltage sources are shifted to those at the two interested reference planes, R_1 and R_2 . To do it, the numerical short-open calibration (SOC) procedure is executed to effectively calibrate out the numerical errors in the MoM as comprehensively discussed in [87]. Next, the J-inverter network parameters in Fig. 8.2(c) can be analytically derived under the equivalence of Y- and J-networks as explained in Chapter 7. The relevant formulas for the coupled line (Type-I) are reported in Chapter 7.2 and those for the coupled line (Type-II) are newly derived and given as below.

As studied for the Type-I structure in Chapter 7.2, the distributed J-susceptance seems to vary as a quasi-periodical function of frequency. In particular, its value achieves the maximal peak at the frequency of $\theta/2 = 90^\circ$ and null at the frequency of $\theta/2 = 180^\circ$. As discussed in [130], the coupling peak in the Type-II structure basically occurs at the frequencies of $\theta/2 = 45^\circ$ and $\theta/2 = 135^\circ$ while the coupling null $\bar{J}$ is placed at the frequency of $\theta/2 = 90^\circ$.

Fig. 8.3(a) and 8.3(b) illustrate the derived normalized J-inverter susceptance ($\bar{J}$) and equivalent electrical length ($\theta/2$) of the two considered coupling structures (Type-I and Type-II) over the frequency range covering the dual-band at 2.4 and 5.2 GHz. For a comparative study, the coupling lengths of these two structures are kept to be identical, i.e. $L_{C1} = L_{C2}$, and they are separately selected as 5.60, 7.30, 9.00 mm in our numerical modeling. The results in Fig. 8.3(a) at first confirm the above-described coupling properties that the two peaks are observed in the type-II

Fig. 8.3. Extracted J-inverter network parameters of the two coupled microstrip lines, namely, Type-I and Type-II. (a) Normalized J-susceptance. (b) Electrical lengths.

structure. Regardless of varied coupled lengths, the 2nd peak in the parameter ($\bar{J}$) is almost located at the frequency of the three times of its 1st peak counterpart. In particular, the coupling null appears around the frequency of $\theta/2 = 90^\circ$ between the 1st and 2nd peaks, and this null can be utilized to achieve the good in-between

isolation between the dual passbands in the dual-band filter design that will be implemented later on. Considering that the coupling degree in the lower band or 2.4GHz-band is always weaker than that in Type-I, this Type-II parallel coupled line seems to be better installed in the middle region between the adjacent resonators in the design of multi-stage dual-band filters. Fig. 8.1(b) shows the schematic of such a filter with the two stages.

In order to achieve the high return losses within the dual passbands and control the dual bandwidths with freedom, the coupling degrees are further comprehensively studied at the central frequencies of the concerned dual bands, i.e. 2.4 and 5.2 GHz. The calculated $\bar{J}$ of the Type-II structure is plotted as a function of the coupling length (L_{C2}) at 2.4 and 5.2 GHz under different coupling spacing (S_2) as illustrated in Fig. 8.4(a) and 8.4(b). In the 2.4GHz-band, $\bar{J}$ seems to go up and then down with respect to L_{C2} , and its maximum gradually rises up as the spacing (S_2) becomes small. Meanwhile, in the 5.2GHz-band, $\bar{J}$ appears relatively weak in the short region of L_{C2} , becomes null at a certain length and then increases quickly to the high peak as L_{C2} is further stretched. In short, the results in Fig. 8.4 provide us with a great capability and flexibility in simultaneously adjusting and controlling the dual-passband response via proper selection of L_{C2} and S_2 .

(a)

(b)

Fig. 8.4. Normalized J-susceptance of the anti-parallel coupled line (Type-II) versus the coupling length (L_{C2}) under different coupled spacing (S_2) at the centers of the dual passbands. (a) $f=2.4$ GHz. (b) $f=5.2$ GHz.

8.3 RESONANT PROPERTIES OF DUAL-BAND RESONATOR

Fig. 8.5(a) and 8.5(b) show the layout and equivalent circuit network of half a symmetrical structure of the two-stage dual-band bandpass filter in Fig. 8.1(b). Fig. 8.5(c) depicts the equivalent cascaded line network of the single stepped-impedance resonator (SIR) [59], [60], [129] that is not ideally isolated with any other line or circuit. Herein, Y_1 and Y_2 are the two unequal characteristic admittances and their aspect ratio can be properly determined to produce the two resonant frequencies at the centers of the specified dual passbands in the filter design. As illustrated in Fig. 8.5(a), the two distinctive coupled lines, i.e. Type-I and Type-II, are capacitively coupled to the single SIR resonator at the two terminals. Following the description given in the above section, the two reference planes ($\#_2$ and $\#_2'$) are selected in Fig. 8.5(a) in order to derive the relevant equivalent network in Fig. 8.5(b). In this aspect, the overall electrical length Φ of this SIR resonator should be made up of the five

Fig. 8.5. Layout and equivalent circuit network of half a symmetrical structure of the proposed two-stage dual-band bandpass filter in Fig. 8.1(b). (a) Layout. (b) Equivalent network. (c) Equivalent SIR resonator.

distinct parts, i.e. $\Phi = \theta_2' + \theta_2'' + \theta_1 + \theta_3'' + \theta_3'$, as shown in Fig. 8.5(b).

Now, we'd like to establish the condition of dual-band resonances at 2.4 and 5.2 GHz with reference to Fig. 8.5(c). Herein, all the phase-related effects caused by the open-end discontinuity and coupled-line are taken into account in the electrical lengths, θ_2' and θ_3' . Thus, the resultant input admittance (Y_{in}) at the left side of this SIR can be easily derived by gradually shifting the input admittance at the different connecting nodes of the two cascaded lines from the right side, such that

$$Y_{in} = jY_2 \frac{R_Z Y^* (1 - \tan \theta_2' \cdot \tan \theta_2'') + \tan \theta_2' + \tan \theta_2''}{1 - \tan \theta_2' \cdot \tan \theta_2'' - R_Z Y^* (\tan \theta_2' + \tan \theta_2'')} \quad (8.1)$$

where,

$$Y^* = \frac{R_Z \tan \theta_1 (1 - \tan \theta_3' \cdot \tan \theta_3'') + \tan \theta_3' + \tan \theta_3''}{R_Z (1 - \tan \theta_3' \cdot \tan \theta_3'') - \tan \theta_1 (\tan \theta_3' + \tan \theta_3'')} \quad (8.2a)$$

$$R_Z = Y_1 / Y_2 \quad (8.2b)$$

By enforcing $Y_{in} = 0$, the multiple resonant frequencies can be numerically solved and the first two ones are actually utilized for dual-band filter design in this work. In addition to the admittance ratio (R_Z), the two resonant frequencies are also affected by the three separate lengths in the SIR.

8.4 DUAL-BAND FILTER

On a basis of the coupling behaviors and dual-band resonances discussed above, the two-stage bandpass filters as shown in Fig. 8.1(b) are initially constructed and designed. In this section, our effort is made to confirm that the dual-band passband response can be really tuned and controlled with freedom. In the design procedure, the SIR resonators are firstly characterized to simultaneously resonate at 2.4 GHz and 5.2 GHz. Secondly, the coupling length (L_{C1}) in the Type-I structure is properly adjusted to achieve the tight and/or weak coupling degrees in the dual passbands with wide and/or narrow bandwidths. Next, the coupling length (L_{C2}) and the gap distance (S_2) in the Type-II structure are determined based on the results in Fig. 8.4 towards maximizing the return losses in the dual passbands. All the optimal design is carried out based on the simple transmission line cascaded network, as shown in Fig. 8.5(b) instead of the time-consuming full-wave simulation over the filter layouts as executed in [59]–[62], [62].

To realize the good in-between isolation of the dual passbands, the transmission zero with $\bar{J}=0$, as shown in Fig. 8.3(a), in the Type-II coupling structure is preferably relocated at the center between the dual passbands at 2.4 and 5.2 GHz. With reference to the design curves in Fig. 8.4, the length (L_{C2}) has to be selected in the following range in the considered case that the coupled spacing varies from 0.2 to 0.6 mm.

$$5.2mm < L_{C2} < 9.4mm \quad (8.3)$$

Following the above discussion, the optimization design is carried out to minimize the return losses in both two passbands with the dual passbands of the proposed dual-band filter in Fig. 8.1(b). Fig. 8.6(a) to 8.6(c) show the optimized dual-band frequency response based on the cascaded transmission line network which parameters are numerically de-embedded from the hybrid method of moments and short-open calibration technique, namely, MoM-SOC. The network-derived S-parameters of all the three dual-band filters are well verified by the direct full-wave Momentum simulation as illustrated in Fig. 8.6. We can see that the return losses within the dual passbands have been maximized above 20dB. The bandwidths of these dual passbands can be really controlled if the coupled lengths/widths (L_{C1} , L_{C2} , S_2) of the Type-I and Type-II coupling structures are coincidentally changed on a basis of the design curves in Fig. 8.4. The fractional bandwidths (FBWs) in the 1st and 2nd passbands (f_0 & f_{s1}) of these filters are tabulated in Table 8.1. This table shows us that the FBW of the 1st passband is significantly reduced from 9.4% to 4.1% while that of the 2nd passband is slightly increased from 3.8% to 5.8%. Thus,

(a)

Fig. 8.6. Predicted results of the three dual-band bandpass filters with varied fractional bandwidths. (a) Type-A. (b) Type-B. (c) Type-C.

the proposed approach is confirmed to have an attractive capability in controlling the dual-band FBWs that is useful for different dual-band communication applications.

Compared to the initial dual-band filters in Chapter 7.2, the proposed dual-band filters here have the excellent in-between isolation of these dual passbands. As shown in Fig. 8.6(a) to 8.6(c), the transmission zero introduced by the Type-II

TABLE 8.1.
MEASURED DUAL-BAND FBWs OF THREE FILTERS

Filter	type-A	type-B	type-C
FBW_{f0}	9.4%	6.6%	4.1%
FBW_{fs1}	3.8%	5.6%	5.8%

coupling structure actually significantly sharpens the in-between stopband, thus effectively suppressing the image signals as requested in the dual-band receiver link design [62].

To achieve the better dual-band performance with deepened in-between stopband, a three-stage dual-band filter is made up by placing an additional resonator in the center between the two resonators used above. Its dual-passband performance is optimally designed using the cascaded transmission line theorem and the predicted S-parameters are plotted in Fig.8.7 together with the full-wave Momentum-

Fig. 8.7. Predicted results of the three-stage dual-band bandpass filter.

simulated results. It can be seen that the return losses in both 2.4GHz- and 5.2GHz-bands are higher than 20 dB. In the meantime, the maximum insertion losses in the in-between stopband exceed 90 dB in the MoM-SOC-derived network simulation

Fig. 8.8. Photograph of the fabricated three-stage dual-band bandpass filter.

Fig. 8.9. Experimental results of the fabricated three-stage dual-band bandpass filter.

and 60 dB in the full-wave Momentum simulation, thereby exhibiting the very good in-between isolation of these dual passbands.

Fig. 8.8 and Fig. 8.9 show the photograph and experimental frequency response of the fabricated three-stage filter circuit. The 3rd resonator in the middle is symmetrically coupled with the 1st and 2nd resonators at the two sides via the Type-II coupling structure. The measured S-parameters show that the stopband between 2.40 and 5.20 GHz passbands achieves the fractional bandwidth of 53% under the 40 dB insertion loss and the maximum insertion loss in this stopband exceeds 65 dB. The minimum insertion losses in the dual passbands achieve 2.44 and 3.52 dB and

the central frequencies of these dual bands are located at 2.42 and 5.19 GHz, respectively.

8.5 CONCLUSION

This work presents a novel dual-band bandpass filter with controllable fractional bandwidths and good in-between isolation in the middle stopband. The frequency-dependent properties of the parallel and anti-parallel coupled microstrip lines are thoroughly investigated to quantitatively demonstrate their distributed coupling performance in terms of extracted J-inverter network parameters and carrying out the efficient optimization design of the whole filter structures based on the simple cascaded network topology. In particular, the parallel coupled line with null coupling at a certain frequency is utilized to produce a preferred transmission zero in the middle stopband, thus effectively isolating these dual passbands. Using these coupling structures and dual-band stepped-impedance resonators, the three dual-band filters with two SIRs are designed and fabricated to exhibit that the dual-passband fractional bandwidths can be really tuned to a large extent. Finally, a three-stage dual-band filter is designed, fabricated and measured to show the improved dual-band filtering performances with sharpened rejection skirt outside the dual passbands and deepened in-between stopband.

Chapter 9

Conclusions and Recommendations

9.1 CONCLUSIONS

In this dissertation, a modified microstrip line with the finite-width ground plane is investigated to construct the potential high-quality RF and microwave integrated circuits. The generalized finite-ground microstrip line (FGMSL) open-end discontinuities with different configurations are characterized by the self-calibrated MoM-SOC technique and the extracted results are provided to derive its dynamic circuit model parameters without any hypothesis in theory. It is exhibited here that the FGMSL open-end discontinuity can still be modeled as a simple static fringing capacitance and its value increases as the finite-ground dimension is enlarged. Then, a novel periodic electromagnetic bandgap structure is presented, characterized and implemented. Using the high-impedance property of the offset FGMSL, the bandwidth and attenuation depth of the concerned guided-wave bandgap are quantitatively confirmed to gain the significant enhancement. The two effective per-unit-length parameters are derived to display the fundamental guided-wave

characteristics of the FGMSL EBG with infinite-extended length. The scattering parameters of the finite-cell EBG circuits with varied feed-line impedances are further investigated to exhibit the bandstop behaviours as a two-port filtering circuit. Subsequently, the offset FGMSL with narrow strip/ground width is studied to show its enhanced characteristic impedance and applied as an improved series inductive element for harmonic-suppressed lowpass filter design. By forming the shortened offset-FGMSL with high impedance, a new class of lowpass filters with miniaturized-size and enhanced-stopband are then proposed, designed and realized. Their advantageous filter performance is quantitatively exhibited and discussed in comparison with the conventional MSL LPF.

The periodically nonuniform coupled microstrip lines (PNCML) are thoroughly characterized as an equivalent uniform coupled transmission line via self-calibrated MoM-SOC technique. For the first time, both two effective per-unit-length transmission parameters, i.e. phase constants and characteristic impedances, are extracted to demonstrate the frequency-dispersive and periodicity-dependent guided-wave characteristics of the two dominant propagating modes. These results are believed to be useful for accurate design of advanced microstrip couplers and filters with the use of such PNCML structures. As an application, our effort is primarily made to predict the location of transmission zero of the finitely-extended PNCML section and further allocate it to suppress the 1st-harmonic passband in the design of microstrip bandpass filters. The PNCMLs with finite length is studied in terms of a generalized coupling-oriented J-inverter network. It does exhibit the basic principle of suppressing the 1st-harmonic passband in filter design with the use of the transmission zero reallocation technique. Extensive results have demonstrated that

the transmission zero or J-susceptance null of the periodically nonuniform coupled microstrip line can be reallocated to cancel the 1st-harmonic resonance. This is done by adjusting the periodicity and/or loaded slit dimension properly.

Dual-band filters have been recently arousing a booming attention as one of the most important circuit blocks in exploring the advanced dual-band wireless systems. A new dual-band filter with controllable fractional bandwidths is proposed and constructed by using the distributed PCML coupled-line and dual-resonance SIR. After coupling dispersion of the PCML with varied lengths is studied in terms of explicit J-inverter susceptance, the three SIR dual-band filters are modeled to exhibit the two tunable passbands. Further, a novel compact dual-band bandpass filter is proposed using the sinuous dual-resonance SIR and dual-band parallel-coupled lines. Without needing any external dual-band impedance-transformer feeds, this filter allows us not only to significantly miniaturize the overall size, but also to achieve the good dual passband responses at 2.4 and 5.2 GHz. Finally, a novel dual-band bandpass filter is presented with controllable fractional bandwidths and good in-between isolation in the middle stopband. The frequency-dependent properties of the pro- and anti-parallel coupled microstrip lines are thoroughly investigated to originally demonstrate their distributed coupling performance and employ the extracted J-inverter network parameters for efficient network-oriented optimization design of the whole filter structures. In particular, the anti-parallel coupled line with null coupling at a certain frequency is utilized to produce a preferred transmission zero in the middle stopband, thus effectively isolating these dual passbands. Using these coupling structures and dual-band stepped-impedance resonators, the three

dual-band filters with two SIRs are designed and fabricated to exhibit that the dual-passband fractional bandwidths can be tuned and controlled to a large extent.

9.2 FUTURE RECOMMENDATIONS

- Microstrip antennas usually have high input impedance at the resonant frequency. Due to the high-impedance property as exhibited here, the offset FGMSL EBG structure is a promising structure to match the microstrip antennas with the feed line in impedance.
- The other FGMSL discontinuities and structures, such as finite-ground microstrip gap and right-angle bend, could be further investigated for establishing a useful library of FGMSL-based circuit elements. Similar to the CPS discontinuities [152], FGMSL series gap can be made up, which consists of an air gap along one of the strip conductors while the other line is kept intact. As derived in [88], the built series capacitance and inductance are put together to make up an equivalent series resonant circuit, where any potential radiation losses can be taken into account by a shunt conductance and a series resistance. In this way, this vivid behavior of radiating resonance may be utilized in the novel design of the advanced microwave antennas.
- The coupling behaviors of PNCML (A) can be further used to make up an advanced microstrip line couplers with high directivity and wide bandwidth.

- In addition, the PNCML can be investigated with respect to varied finite-ground plane widths and configurations so as to provide the more degree of freedom in the design of advanced microwave integrated circuits. As a trade-off of PNCML structures, the coupling between the two nonuniform strip lines is decreased as the depth of periodically transverse slits is increased. This drawback can be compensated by unitizing the advantages of the FGMSL, which is helpful to increase the coupling of PNCML, also in size compaction.
- As discussed in Chapter 8, the coupling null of pro-parallel coupled line appears at the frequency with the electrical length of $\theta/2 = 90^\circ$ that is sandwiched between the 1st and 2nd peaks. This null characteristic may be further applied for parallel coupled line filters with harmonic suppression.

Author's Publications

PUBLISHED JOURNAL PAPERS

- [1] **S. Sun** and L. Zhu, "Novel design of microstrip bandpass filters with a controllable dual-passband response: description and implementation," *IEICE Trans. Electron.*, vol. E89-C, no. 2, pp. 197-202, Feb. 2006. (The contents of Chapter 8)
- [2] L. Zhu, **S. Sun** and W. Menzel, "Ultra-wide band (UWB) bandpass filters using multiple-mode resonator," *IEEE Microwave Wireless Component Lett.*, vol. 15, no. 11, pp. 796-798, Nov. 2005.
- [3] **S. Sun** and L. Zhu, "Electromagnetic bandgap enhancement using the high-impedance property of offset finite-ground microstrip line," *Microwave Optical Technology Lett.*, vol. 47, issue. 6, pp. 543-546, 20 Dec. 2005, Published Online: 25 Oct. 2005. (The contents of Chapter 3, Section 3.3)
- [4] **S. Sun** and L. Zhu, "Compact dual-band microstrip bandpass filter without external feeds," *IEEE Microwave Wireless Component Lett.*, vol. 15, no. 10, pp. 644-646, Oct. 2005. (The contents of Chapter 7, Section 7.3)
- [5] **S. Sun** and L. Zhu, "Stopband-enhanced and size-miniaturized lowpass filters using high-impedance property of offset finite-ground microstrip line," *IEEE Trans. Microwave Theory Tech.*, vol. 53, no. 9, pp. 2844-2850, Sept. 2005. (The contents of Chapter 4)
- [6] **S. Sun** and L. Zhu, "Periodically nonuniform coupled microstrip lines with equalized even- and odd mode velocities for harmonic suppression in filter design," *IEICE Trans. Comm.*, vol. E88-B, no. 6, pp. 2377-2382, Jun. 2005. (The contents of Chapter 6, Section 6.2)
- [7] **S. Sun** and L. Zhu, "Design of periodically nonuniform coupled microstrip filters with harmonic suppression using equivalent J-inverter network," *IEEE Trans. Microwave Theory Tech.*, vol. 53, no. 5, pp. 1817-1822, May 2005. (The contents of Chapter 6, Section 6.3)
- [8] **S. Sun** and L. Zhu, "Guided-wave characteristics of periodically nonuniform coupled microstrip lines: even and odd modes," *IEEE Trans. Microwave Theory Tech.*, vol. 53, no. 4, pp. 1221-1227, Apr. 2005. (The contents of Chapter 5)
- [9] **S. Sun** and L. Zhu, "Unified equivalent circuit model of finite-ground microstrip line open-end discontinuities using MoM-SOC technique," *IEICE*

Trans. Electron., vol. E87-C, no. 5, pp. 828-831, May 2004. (The contents of Chapter 3, Section 3.2)

PUBLISHED CONFERENCE PAPERS

- [1] **S. Sun** and L. Zhu, "Novel design of microstrip bandpass filters with a controllable dual-passband response: description and implementation", in *Asia Pacific Microwave Conf.*, Dec. 2005, pp. 729-732. (The contents of Chapter 8, Section 8.4)
- [2] **S. Sun** and L. Zhu, "Coupling dispersion of parallel-coupled microstrip lines for dual-band filters with controllable fractional pass bandwidths," in *IEEE MTT-S Int. Microwave Symp.*, June 2005, pp. 2195-2198. (The contents of Chapter 7, Section 7.2)
- [3] **S. Sun** and L. Zhu, "Periodic finite-ground microstrip line with high-impedance offset sections: enhanced electromagnetic bandgap," in *IEEE Int. Workshop on Antenna Tech. Small Antennas and Novel Metamaterials (IWAT)*, March 2005, pp. 383-386. (The contents of Chapter 3, Section 3.3)
- [4] **S. Sun** and L. Zhu, "Guided-wave characteristics of periodically nonuniform coupled microstrip lines for harmonic suppression in filter design," in *IEICE Int. Symp. on Antennas and Propagation*, August 2004, pp. 449-452. (The contents of Chapter 6, Section 6.2)

Bibliography

- [1] A. A. Oliner, "Historical perspectives on microwave field theory," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, no. 9, pp. 1022–1045, Sept. 1984.
- [2] H. Sobol and K. Tomiyasu, "Milestones of microwaves," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 3, pp. 594–611, Mar. 2002.
- [3] J. H. Howe, "Microwave integrated circuits – an historical perspective," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, no. 9, pp. 991–996, Sept. 1984.
- [4] J. D. N. Mcquiddy, J. W. Wassel, J. B. Lagrange and W. Wisseman, "Monolithic microwave integrated circuits – an historical perspective," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, no. 9, pp. 997–1008, Sept. 1984.
- [5] E. C. Niehenke, R. A. Pucel and I. J. Bahl, "Microwave and millimeter-wave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 3, pp. 846–857, Mar. 2002.
- [6] K. C. Gupta, R. Garg, I. Bahl and P. Bhartia, *Microstrip Lines and Slotlines*, 2nd ed. Norwood, MA: Artech House, 1996.
- [7] T. Itoh, *Planar Transmission Line Structures*, New York: IEEE Press, 1987.
- [8] R. N. Simons, *Coplanar Waveguide Circuits, Components, and Systems*, John Wiley & Sons, 2001.
- [9] D. D. Grieg and H. F. Engelmann, "Microstrip – A new transmission technique for the kilomegacycle range," *Proc. IRE*, vol.40, pp. 1644-1650, Dec. 1952.
- [10] F. Assadourian and E. Rimai, "Simplified theory of microstrip transmission systems," *Proc. IRE*, vol.40, pp. 1651-1657, Dec. 1952.
- [11] J. A. Kostriza, "Microstrip components," *Proc. IRE*, vol.40, pp. 1658-1663, Dec. 1952.
- [12] R. E. Collin, *Foundations for microwave engineering*, 2nd edition, McGraw Hill, 1992.
- [13] D. M. Pozar, *Microwave Engineering*, 2nd edition, John Wiley & Sons, 1998.
- [14] C. A. Balanis, *Antenna theory - analysis and design*, 2nd edition, John Wiley & Sons, 1996.

-
- [15] C. E. Smith and R. -S. Chang, "Microstrip transmission line with finite-width dielectric", *IEEE Trans. Microwave Theory and Tech.*, vol. MTT-28, pp. 90-94, Feb. 1980.
- [16] C. E. Smith and R. -S. Chang, "Microstrip transmission line with finite-width dielectric and ground plane", *IEEE Trans. Microwave Theory and Tech.*, vol. MTT-33, pp. 835-839, Sept. 1985.
- [17] J. Venkataraman, S. M. Rao, A. R. Djordjevic, T. K. Sarkar and Yang Naiheng "Analysis of arbitrarily oriented microstrip transmission lines in arbitrarily shaped dielectric media over a finite ground plane," *IEEE Trans. Microwave Theory and Tech.*, vol. 33, pp. 952-960, Oct. 1985.
- [18] K. Y. See and E. M. Freeman, "Rigorous approach to modeling electromagnetic radiation from finite-size printed circuit structures," *IEE Proc. Microw. Antennas Propag.*, vol. 146, No. 1, pp. 29-34, Feb. 1998.
- [19] L. Gurel and W. C. Chew, "Guidance or resonance conditions for strip or disks embedded in homogeneous and layered media," *IEEE Trans. Microwave Theory and Tech.*, vol. 36, pp. 1498-1506, Nov. 1988.
- [20] G. E. Ponchak, A. Margomenos and Linda P. B. Katehi, "Low loss finite width ground plane, thin film microstrip lines on Si wafers," *Silicon Monolithic Integrated Circuits in RF Systems, Dig.* 2000, pp. 43-47.
- [21] G. E. Ponchak, E. Dalton, E. M. Tentzeris and J. Papapolumerou, "Coupling between microstrip lines with finite width ground plane embedded in polyimide layers for 3D-MMICs on Si," in *IEEE MTT-S Int. Symp. Dig.* vol. 3, 2002, pp. 2221-2224.
- [22] J. Huang, "The finite ground plane effect on the microstrip antenna radiation pattern," *IEEE Trans. Antennas Propag.*, vol. AP-31, No. 4, pp. 649-653, Jul. 1983.
- [23] E. Lier and K. R. Jakobsen, "Rectangular microstrip patch antennas with Infinite and finite ground plane dimensions," *IEEE Trans. Antennas Propag.*, vol. AP-31, No. 6, pp. 978-984, Nov. 1983.
- [24] S. Rengarajan, "Finite ground plane effects on planar arrays of microstrip patch elements," in *IEEE AP-S Int. Symp.*, vol.22, 1984, pp. 741-744.
- [25] E. Lier, "Modified radiation conductance due to finite ground plane of rectangular microstrip antenna," in *IEEE AP-S Int. Symp.*, vol.20, 1982, pp. 281-284.
- [26] A. K. Bhattacharyya, "Effects of finite ground plane on the radiation characteristics of a circular patch antenna," *IEEE Trans. Antennas Propag.*, vol. 38, pp. 152-159, Feb. 1990.

-
- [27] P. Moosavi and L. Shafai, "Directivity of microstrip ring antennas and effects of finite ground plane on the radiation parameters," in *IEEE AP-S Int. Symp.*, vol.2, Jul. 1998, pp. 21-26.
- [28] R. Marg, I. Schon and A.F. Jacob, "Finite ground plane effects on the radiation pattern of small microstrip arrays," *IEE Proc. Microw. Antennas Propag.*, vol. 147, pp. 139-143, Apr. 2000.
- [29] S. Noghanian and L. Shafar, "Control of microstrip antenna radiation characteristics by ground plane size and shape," *IEE Proc. Microw. Antennas Propag.*, vol. 145, No. 3, pp. 207-212, Jun. 1998.
- [30] D. Chatterjee, V. Natarajan, K.-F. Lee and X. Wang, "Techniques for bandwidth optimization of probe-fed microstrip antennas on small, finite ground planes," in *IEEE AP-S Int. Symp.*, vol.3, 2001, pp. 282-285.
- [31] C. C. Lu and C. Yu, "Analysis of microstrip structure of finite ground plane using the hybrid volume-surface integral equation approach," in *IEEE AP-S Int. Symp.*, vol.4, 2002, pp. 162-165.
- [32] D. M. Pozar and N. K. Das, "A generalized spectral-domain green's function for multilayer dielectric substrates with application to multilayer transmission lines," *IEEE Trans. Microwave Theory and Tech.*, vol. 35, pp. 326-335, Mar. 1987.
- [33] W. P. Harokopus and P. B. Katehi, "Characterization of microstrip discontinuities on multilayer dielectric substrates including radiation losses," *IEEE Trans. Microwave Theory and Tech.*, vol. 37, pp. 2058-2066, Dec. 1989.
- [34] H. A. Auda and A. Z. Elsherbeni, "A WH/GSMT-based full-wave analysis for planar transmission lines embedded in multilayered dielectric substrates," *IEEE Trans. Microwave Theory and Tech.*, vol. 43, pp. 119-130, Jan. 1995.
- [35] K. Kamogawa, T. Tokumitsu, M. Aikawa, "A novel microstrip antenna using alumina-ceramic/polyimide multilayer dielectric substrate," in *IEEE MTT-S Int. Symp. Dig.* vol. 1, Jun. 1996, pp. 71-74.
- [36] Y. Qian and E. Yamashita, "A 60 GHz imaging array using CPW-fed twin-slots on multilayered substrates," in *IEEE MTT-S Int. Symp. Dig.* vol. 2, 17-21 Jun. 1996, pp. 1007-1010.
- [37] J.-M. Laheurte, L.P.B. Katehi, G.M. Rebeiz, "CPW-fed slot antennas on multilayer dielectric substrates," *IEEE Trans. Antennas Propagat.*, vol. 44, pp. 1102-1111, Aug. 1996.
- [38] K. Kamogawa, T. Tokumitsu, M. Aikawa, "Multifrequency microstrip antennas using alumina-ceramic/polyimide multilayer dielectric substrate," *IEEE Trans. Microwave Theory and Tech.*, vol. 44, pp. 2431-2437, Dec. 1996.

-
- [39] K. Li, K. Atsuki, "An analysis of traveling-wave optical modulator with thick, trapezoidal electrodes on multilayered dielectric substrate," in *MWP. Tech. Dig., Int.*, 3-5 Dec 1996.
 - [40] V. Losada, R. R. Boix, M. Horno, "Resonant modes of circular microstrip patches in multilayered substrates," *IEEE Trans. Microwave Theory and Tech.*, vol. 47, pp. 488-498, Apr. 1999.
 - [41] J. F. Carpentier, F. Huret, P. Pribetich, P. Kennis, "Comparison between a quasistatic space-domain analysis and a full-wave spectral-domain analysis for high-frequency interconnecting crossover in multilayered dielectric media," *IEE Proc. Microw. Antennas Propag.*, vol. 142, pp. 275 -281, Jun. 1995.
 - [42] K. Lim, S. Pinel, M. Davis, A. Sutono, C. Lee, D. Heo, A. Obatoynbo, J. Laskar, E.M. Tantzaris, R. Tummala, "RF-System-on-package (SOP) for wireless communications," *IEEE Microwave magazine*, pp. 88-99, Mar. 2002.
 - [43] S. Chakraborty, K. Lim, A. Sutono, E. Chen, S. Yoo, A. Obatoyinbo, S.-W. Yoon, M. Maeng, M.F. Davis, S. Pinel, J. Laskar, "A 2.4-GHz radio front end in RF system-on-package technology," *IEEE Microwave magazine*, pp. 94-104, Jun. 2002.
 - [44] L. K. Yeung and K.L. Wu, "A compact second-order LTCC bandpass filter with two finite transmission zeros," *IEEE Trans. Microwave Theory and Tech.*, vol. 51, pp. 337-341, Feb. 2003.
 - [45] S. B. Cohn, "Parallel-coupled transmission-line-resonator filters," *IRE Trans. Microwave Theory Tech.*, vol. MTT-6, pp. 223-231, Apr. 1958.
 - [46] A. Podell, "A high directivity microstrip coupler technique," in *IEEE MTT-S Int. Symp. Dig.*, pp. 33-36, 1970.
 - [47] J.-T. Kuo, W.-H. Hsu and W.-T. Huang, "Parallel coupled microstrip filters with suppression of harmonic response," *IEEE Microwave Wireless Component Lett.*, vol. 12, no.10, pp. 383-385, Oct. 2002.
 - [48] A. Riddle, "High performance parallel coupled microstrip filters," *1988 IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 1, pp. 427-430, 1988.
 - [49] T. Lopetegi, M. A. G. Laso, J. Hernández, M. Bacaicoa, D. Benito, M. J. Garde, M. Sorolla and M. Guglielmi, "New microstrip 'wiggly-line' filters with spurious passband suppression," *IEEE Trans. Microwave Theory Tech.*, vol. 49, no. 9, pp. 1593-1598, Sept. 2001.
 - [50] B. S. Kim, J. W. Lee, and M. S. Song, "Modified microstrip filters improving the suppression performance of harmonic signals," *2003 IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 1, pp. 539-542, June 2003.

-
- [51] P. Vincent, J. Culver and S. Eason, "Meandered line microstrip filter with suppression of harmonic passband response," *2003 IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 3, pp. 1905-1908, June 2003.
 - [52] J.-T. Kuo, M. Jiang and H.-J. Chang, "Design of parallel-coupled microstrip filters with suppression of spurious resonances using substrate suspension," *IEEE Trans. Microwave Theory Tech.*, vol. 52, no. 1, pp. 83-89, Jan. 2004.
 - [53] J.-T. Kuo and M. Jiang, "Enhanced Microstrip Filter Design With a Uniform Dielectric Overlay for Suppressing the Second Harmonic Response," *IEEE Microwave Wireless Component Lett.*, vol. 14, no.9, pp. 419-421, Sept. 2004.
 - [54] M. C. Velazquez-Ahumada, J. Martel and F. Medina, "Parallel coupled microstrip filters with ground-plane aperture for spurious band suppression and enhanced coupling," *IEEE Trans. Microwave Theory Tech.*, vol. 52, no. 3, pp. 1082-1086, Mar. 2004.
 - [55] H. Hashemi and A. Hajimiri, "Concurrent multiband low-noise amplifiers-theory, design and applications," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 1, pp. 288-301, Jan. 2002.
 - [56] F. V. Straten, B. Smolders, A. v. Zuijlen and R. Ooijman, "Multiband cellular RF solutions," *IEEE Journal of Solid-State Circuits*, vol. 39, no. 10, pp. 1598-1604, Oct. 2004.
 - [57] H. Miyake, et al., "A miniaturized monolithic dual band filter using ceramic lamination technique for dual mode portable telephones," *1997 IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 2, pp. 789-792, June 1997.
 - [58] C. H. Chang, H. S. Wu, H. J. Yang and C. K. C. Tzuang, "Coalesced single-input single-output dual-band filter," *2003 IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 1, pp. 511-514, June 2003.
 - [59] S. -F. Chang, J. -L. Chen and S. -C. Chang, "New dual-band bandpass filters with step-impedance resonators comb and hairpin structures," *2003 Proc. Asia Pacific Microwave Conf.*, pp. 793-796, 2003.
 - [60] H. -M. Lee, C. -R. Chen, C. -C. Tsai and C. -M. Tsai, "Dual-band coupling and feed structure for microstrip filter design," *2004 IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 3, pp. 1971-1974, June 2004.
 - [61] J. -T. Kuo and H. -S. Cheng, "Design of quasi-elliptic function filters with a dual-passband response," *IEEE Microwave Wireless Components Lett.*, vol.14, no.10, pp. 472 - 474, Oct. 2004.
 - [62] C. -C. Chen, "Dual-band bandpass filter using coupled resonator pairs," *IEEE Microwave Wireless Components Lett.*, vol.15, no.4, Apr. 2005.

-
- [63] S. M. Ali and S. F. Mahmoud, "Electromagnetic fields of buried sources in stratified anisotropic media," *IEEE Trans. Antennas Propagat.*, vol. AP-27, pp. 671–678, Sept. 1979.
- [64] J. S. Bagby, D. P. Nyquist, and B. C. Drachman, "Integral formulation for analysis of integrated dielectric waveguides," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 906–915, Oct. 1985.
- [65] C. M. Krowne, "Determination of the Green's function in the spectral domain using a matrix method: Application to radiators immersed in a complex anisotropic layered medium," *IEEE Trans. Antennas Propagat.*, vol. AP-34, pp. 247–253, Feb. 1986.
- [66] N. K. Das and D. M. Pozar, "A generalized spectral-domain Green's function for multilayer dielectric substrates with application to multilayer transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-35, pp. 326–335, Mar. 1987.
- [67] L. Beyne and D. De Zutter, "Green's function for layered lossy media with special application to microstrip antennas," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 875–881, May 1988.
- [68] L. Vegni, R. Cicchetti, and P. Capece, "Spectral dyadic Green's function formulation for planar integrated structures," *IEEE Trans. Antennas Propagat.*, vol. 36, pp. 1057–1065, Aug. 1988.
- [69] R. Kastner, E. Heyman, and A. Sabban, "Spectral domain iterative analysis of single- and double-layered microstrip antennas using the conjugate gradient algorithm," *IEEE Trans. Antennas Propagat.*, vol. 36, pp. 1204–1212, Sept. 1988.
- [70] S. Barkeshli and P. H. Pathak, "On the dyadic Green's function for a planar multilayered dielectric/magnetic media," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 128–142, Jan. 1992.
- [71] P. Bernardi and R. Cicchetti, "Dyadic Green's functions for conductor backed layered structures excited by arbitrary tridimensional sources," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 1474–1483, Aug. 1994.
- [72] S.-G. Pan and I. Wolff, "Scalarization of dyadic spectral Green's functions and network formalism for three-dimensional full-wave analysis of planar lines and antennas," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 2118–2127, Nov. 1994.
- [73] A. Dreher, "A new approach to dyadic Green's function in spectral domain," *IEEE Trans. Antennas Propagat.*, vol. 43, pp. 1297–1302, Nov. 1995.
- [74] K. A. Michalski and J. R. Mosig, "Multilayered media Green's functions in integral equation formulations," *IEEE Trans. Antennas Propagat.*, vol. 45, pp. 508–519, Mar. 1997.

-
- [75] T. Itoh, "Spectral Domain Immittance Approach for Dispersion Characteristics of Generalized Printed Transmission Lines," *IEEE Trans. Microwave Theory Tech.*, vol. 28, pp. 733-736, Jul. 1980.
- [76] R. F. Harrington, *Field Computation by Moment Methods*. New York: Macmillan, 1968; Melbourne, FL: Krieger, reprint 1982.
- [77] M. N. O. Sadiku, *Numerical techniques in electromagnetics*, 2nd edition, CRC Press, 2002.
- [78] Cam N., *Analysis methods for RF, microwave, and millimeter-wave planar transmission Line structures*, John Wiley & Sons, 2000.
- [79] P. B. Katehi and N. G. Alexopoulos, "Frequency-dependent characteristics of microstrip discontinuities in millimeter-wave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 1029-1035, Oct. 1985.
- [80] R. W. Jackson and D. M. Pozar, "Full-wave analysis of microstrip open-end and gap discontinuities," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 1036-1042, Oct. 1985.
- [81] R. H. Jansen, "The spectral-domain approach for microwave integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 1043-1056, Oct. 1985.
- [82] J. R. Rautio and R. F. Harrington, "An electromagnetic time-harmonic analysis of shielded microstrip circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-35, pp. 726-730, Aug. 1987.
- [83] L. P. Dunleavy and P. B. Katehi, "A generalized method for analyzing thin microstrip discontinuities," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 1758-1766, Dec. 1988.
- [84] J. R. Mosig, "Arbitrarily shaped microstrip structures and their analysis with a mixed potential integral equation," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 314-323, Feb. 1988.
- [85] M. Kahrizi, T. K. Sarkar, and Z. A. Maricevic, "Analysis of a wide radiation slot in the ground plane of a microstrip line," *IEEE Trans. Microwave Theory Tech.*, vol. 41, pp. 29-37, Jan. 1993.
- [86] L. Zhu and K. Wu, "Characterization of unbounded multiport microstrip passive circuits using an explicit network-based method of moments," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 2114-2124, Dec. 1997.
- [87] L. Zhu and K. Wu, "Unified equivalent-circuit model of planar discontinuities suitable for field theory-based CAD and optimization of M(H)MIC's," *IEEE Trans. Microwave Theory Tech.*, pt. I, vol. 47, pp. 1589 - 1602, Sept. 1999.

-
- [88] L. Zhu and K. Wu, "Field-extracted lumped-element models of coplanar stripline circuits and discontinuities for accurate radiofrequency design and optimization," *IEEE Trans. Microwave Theory Tech.*, vol. 50, pp.1207-1215, Apr. 2002.
- [89] L. Zhu and K. Wu, "Short-open calibration technique for field theory-based parameter extraction of lumped elements of planar integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 50, pp. 1861 -1869, Aug. 2002.
- [90] M. J. Tsai, F. D. Flacis, O. Fordham, N. G. Alexopoulos, "Modeling planar arbitrarily shaped microstrip elements in multilayered media," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 330-337, Mar. 1997.
- [91] N. G. Alexopoulos and S. C. Wu, "Frequency-independent equivalent circuit model for microstrip open-end and gap discontinuities," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-42, no.7, pp. 1268-1272, July. 1994.
- [92] J. S. Kim and W. S. Park, "Efficient analysis of microstrip lines including edge singularities in spatial domains," *IEEE Microwave Wireless Components Lett.*, vol. 11, pp. 270-272, Jun. 2001.
- [93] M. Farina and T. Rozzi, "A short-open deembedding technique for method-of-moments-based electromagnetic analyses," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-49, no.4, pp.624-628, Apr. 2001.
- [94] F. Yang, R. Coccioli, Y. Qian, and T. Itoh, "Planar PBG structures: basic properties and applications," *IEICE Trans. Electron.*, vol. E83-C, no. 5, pp. 687-696, May. 2000.
- [95] D. Dawn, Y. Ohashi and T. Shimura, "A novel electromagnetic bandgap metal plate for parallel plate mode suppression in shielded structures," *IEEE Microwave Wireless Components Lett.*, vol.12, no.5, pp.166-168, May. 2002.
- [96] R. Abhari and G. V. Eleftheriades, "Metallo-dielectric electromagnetic bandgap structures for suppression and isolation of the parallel-plate noise in high-speed circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 51, no. 6, pp. 1629-1639, Jun. 2003.
- [97] C. -K. Wu, H. -S. Wu and C. -K. C. Tzuang, "Electric-magnetic-electric slow-wave microstrip line and bandpass filter of compressed size," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 8, pp. 1996-2004, Aug. 2002.
- [98] S. -G. Mao and M. -Y. Chen, "Propagation characteristics of finite-width conductor-backed coplanar waveguides with periodic electromagnetic bandgap cells," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 11, pp. 2624-2628, Nov. 2002.
- [99] L. Zhu, "Guided-wave characteristics of periodic microstrip lines with inductive loading: slow-wave and bandstop behaviors," *Microwave and Optical Technology Lett.*, vol. 41, no. 2, pp. 77-79, Apr. 2004.

-
- [100] S. Sun and L. Zhu, "Unified equivalent circuit model of finite-ground microstrip line open-end discontinuities using MoM-SOC technique," *IEICE Trans. Electron.*, vol. E87-C, no. 5, pp. 828-831, May. 2004.
 - [101] L. Zhu, "Guided-wave characteristics of periodic coplanar waveguides with inductive loading – unit-length transmission parameters," *IEEE Trans Microwave Theory Tech.*, vol.51, no.10, pp.2133-2138, Oct. 2003.
 - [102] L. Zhu and K. Wu, "Revisiting characteristic impedance and its definition of microstrip line with a self-calibrated 3-D MoM scheme," *IEEE Microwave Wireless Components Lett.*, vol.8, no.2, pp.87-89, Feb. 1998.
 - [103] G. L. Matthaei, L. Young and E. M. T. Jones, *Microwave filters, impedance-matching networks, and coupling structures*, Dedham, MA: Artech house, 1980.
 - [104] J. S. Hong and M. J. Lancaster, *Microstrip filters for RF/microwave applications*, John Wiley & Sons, Inc, 2001.
 - [105] L. Zhu, H. Bu, and K. Wu, "Unified CAD model of microstrip line with back side aperture for multilayer integrated circuit," in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol.2, pp.981-984, June 2000.
 - [106] D. Ahn, J. S. Park, C. S. Kim, J. Kim, Y. Qian and T. Itoh, "A design of the low-pass filter using the novel microstrip defected ground structure," *IEEE Trans. Microwave Theory Tech.*, vol. 49, no. 1, pp. 86-93, Jan. 2001.
 - [107] L. H. Hsieh and K. Chang, "Compact elliptic-function low-pass filters using microstrip stepped-impedance hairpin resonators," *IEEE Trans. Microwave Theory Tech.*, vol. 51, no. 1, pp. 193-199, Jan. 2003.
 - [108] K. Kim, S. Kim, H. Han, I. park and H. Lim, "Compact microstrip lowpass filter using shunt open stubs and coupled slots on ground plane," *IEE Electronics Lett.*, vol. 40, no. 5, pp. 313-314, 4 Mar. 2004.
 - [109] S. G. Kim and K. Chang, "Ultrawide-band transitions and new microwave components using double-sided parallel-strip lines", *IEEE Trans. Microwave Theory Tech.*, vol. 52, no. 9, pp. 2148-2152, Sept. 2004.
 - [110] T. Sugiura, "Analysis of distributed-lumped strip transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. 25, no. 8, pp. 656-661, Aug. 1977.
 - [111] F. -J. Glandorf and I. Wolff, "A spectral-domain analysis of periodically nonuniform coupled microstrip lines," *IEEE Trans. Microwave Theory Tech.*, vol. 36, no. 3, pp. 522-528, Mar. 1988.
 - [112] J. -N. Hwang and J.-T. Kuo, "FDTD analysis of periodically nonuniform coupled microstrip lines," *2003 IEEE AP-S Int. Symp. Dig.*, vol. 1, pp. 741-744, June 2003.

-
- [113] V. I. Okhmatovski, J. Morsey and A. C. Cangellaris, "On deembedding of port discontinuities in full-wave CAD models of multiport circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 51, no. 12, pp. 2355-2365, Dec. 2003.
- [114] L. Zhu, "Unified 3-D definition of CPW- and CSL-mode characteristic impedances of coplanar waveguide using MoM-SOC technique," *IEEE Microwave Wireless Components Lett.*, vol.13, no.4, pp.158-160, April 2003.
- [115] Advanced Design System 2003a, provided by Agilent Technologies, CA, U.S.A.
- [116] S. Denis, C. Person, S. Toutain, B. Théron and S. Vigneron, "Parallel coupled microstrip filter with phase difference compensation," *Electron. Lett.*, vol. 31, no. 22, pp. 1927-1928, 26 Oct. 1995.
- [117] S.-F. Chang, Y.-H. Jeng and J.-L. Chen, "Tapped wiggly-coupled technique applied to microstrip bandpass filters for multi-octave spurious suppression," *Electron. Lett.*, vol. 40, no. 1, pp.46-47, 8 Jan. 2004.
- [118] J. García-García, F. Martín, F. Falcone, J. Bonache, I. Gil, T. Lopetegi, M. A. G. Laso, M. Sorolla and R. Marqués, "Spurious passband suppression in microstrip coupled line band pass filters by means of split ring resonators," *IEEE Microwave Wireless Component Lett.*, vol. 14, no. 9, pp. 416- 418, Sept. 2004.
- [119] L. Zhu, H. Bu and K. Wu, "Broadband and compact multi-pole microstrip bandpass filters using ground plane aperture technique" *IEE Proc. Microw. Antennas Propag.*, vol. 149, no. 1, pp. 71-77, Feb. 2002.
- [120] G. I. Zysman and A. K. Johnson, "Coupled transmission line networks in an inhomogeneous dielectric medium," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-17, no. 10, pp. 753-759, Oct. 1969.
- [121] M. C. Velazquez-Ahumada, J. Martel and F. Medina, "Parallel coupled microstrip filters with floating ground-plane conductor for spurious-band suppression," *IEEE Trans. Microwave Theory Tech.*, vol. 53, no. 5, pp. 1823-1828, May 2005.
- [122] J. Kuo, S. Chen and M. Jiang, "Parallel coupled microstrip filters with over-coupled end stages for suppression of spurious responses," *IEEE Microwave Wireless Component Lett.*, vol. 13, no. 10, pp. 440- 442, Oct. 2003.
- [123] P. Cheong, S. W. Fok and K. W. Tam, "Miniaturized parallel coupled-line bandpass filter with spurious-response suppression," *IEEE Trans. Microwave Theory Tech.*, vol. 53, no. 5, pp. 1810-1816, May 2005.
- [124] P. Cheong, S. W. Fok and K. W. Tam, "Miniaturized spurious passband suppression microstrip filter using meandered parallel coupled lines," *IEEE Trans. Microwave Theory Tech.*, vol. 53, no. 2, pp. 747-753, Feb. 2005.

-
- [125] "Transmission Line Calculator", Applied Wave Research, Inc. 2001.
 - [126] C. Quendo, E. Rius and C. Person, "An original topology of dual-band filter with transmission zeros," *2003 IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 2, pp. 1093-1096, June 2003.
 - [127] L. -C. Tsai and C. -W. Hsue, "Dual-band bandpass filters using equal-length coupled-serial-shunted lines and Z-transform technique," *IEEE Trans. Microwave Theory Tech.*, vol. 52, no. 4, pp. 1111-1117, Apr. 2004.
 - [128] M. Makimoto and S. Yamashita, "Bandpass filters using parallel coupled stripline stepped impedance resonators," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-28, no. 12, pp. 1413-1417, Dec. 1980.
 - [129] S. -F. Chang, W. -L. chen, S. -C. Chang, C. -K. Tu, C. -L. Wei, C. -H. Chien, C. -H. Tsai, J. Chen and A. Chen, "A dual-band RF transceiver for multistandard WLAN applications," *IEEE Trans. Microwave Theory Tech.*, vol. 53, no. 3, pp. 1048-1055, Mar. 2005.
 - [130] M. Matsuo, H. Yabuki and M. Makimoto, "Improvement of stop-band characteristics for half-wavelength resonator filters," *IEICE Trans. Electron.*, vol. E85-C, no. 7, pp. 1472-1477, Jul. 2002.
 - [131] T. Lopetegi, M. A. G. Laso, F. Falcone, F. Martin, J. Bonache, L. P. Cuevas and M. Sorolla, "Microstrip wiggly line bandpass filters with multispurious rejection," *IEEE Microwave Wireless Component Lett.*, vol. 14, no. 11, pp. 531- 533, Nov. 2004.
 - [132] F. R. Yang, K. P. Ma and T. Itoh, "A uniplanar compact photonic-bandgap (UC-PBG) structure and its applications for microwave circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 47, no. 8, pp. 1509-1514, Aug. 1999.
 - [133] C. K. Wu, H. S. Wu and C. K. C. Tzuang, "Electric-magnetic-electric slow-wave microstrip line and bandpass filter of compressed size," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no. 8, pp. 1996-2004, Aug. 2002.
 - [134] J. García-García, F. Martín, F. Falcone, J. Bonache, J. D. Baena, I. Gil, E. Amat, T. Lopetegi, M. A. G. Laso, J. A. M. Iturmendi, M. Sorolla and R. Marqués, "Microwave filters with improved stopband based on sub-wavelength resonators," *IEEE Trans. Microwave Theory Tech.*, vol. 53, no. 6, pp. 1997-2006, Jun. 2005.
 - [135] K. Wada, K. Nakagawa, O. Hashimoto and H. Harada, "Technique for improving out-of-band characteristics of planar microwave filters using tapped resonators," *IEICE Trans. Electron.*, vol. E85-C, no. 2, pp. 391-399, Feb. 2002.
 - [136] K. Wada, T. Ohno, K. Nakagawa and O. Hashimoto, "Realization of low spurious responses by various bandpass filters using open-ended $\lambda/2$ resonators," *IEICE Trans. Electron.*, vol. E86-C, no. 12, pp. 2394-2402, Dec. 2003.

-
- [137] C. S. Lee, V. Nalbandian and F. Schwering, "Planar dual-band microstrip antenna," *IEEE Trans. Antennas Propagat.*, vol. 43, no. 8, pp. 892-894, Aug. 1995
 - [138] F. Tefiku, ber and C. A. Grimes, "Design of broad-band and dual-band antennas comprised of series-fed printed-strip dipole pairs," *IEEE Trans. Antennas Propagat.*, vol. 48, no. 6, pp. 895-900, Jun. 2000.
 - [139] Y. L. Kuo, K. L. Wong, "Printed double-T monopole antenna for 2.4/5.2 GHz dual-band WLAN operations," *IEEE Trans. Antennas Propagat.*, vol. 51, no. 9, pp. 2187-2192, Sep. 2003.
 - [140] Y. -X. Guo, M. Y. W. Chia and Z. N. Chen, "Miniature built-in multiband antennas for mobile handsets," *IEEE Trans. Antennas Propagat.*, vol. 52, no. 8, pp. 1936-1944, Aug. 2004.
 - [141] K. L. Wong, L. C. Chou and C. M. Su, "Dual-band flat-plate antenna with a shorted parasitic element for laptop applications," *IEEE Trans. Antennas Propagat.*, vol. 53, no. 1, pp. 539-544, Jan. 2005.
 - [142] Z. Zhang, M. F. Iskander, J. C. Langer, J. Mathews, "Dual-band WLAN dipole antenna using an internal matching circuit," *IEEE Trans. Antennas Propagat.*, vol. 53, no. 5, pp. 1813-1818, May. 2005.
 - [143] A. Schmidt and S. Catala, "A universal dual band LNA implementation in SiGe technology for wireless applications," *IEEE Journal of Solid-State Circuits*, vol. 36, no. 7, pp. 1127-1131, Jul. 2001.
 - [144] S. F. R. Chang, W. L. Chen, S. C. Chang, C. K. Tu, C. L. Wei, C. H. Chien, C. H. Tsai, J. Chen and A. Chen, "A dual-band RF transceiver for multistandard WLAN applications," *IEEE Trans. Microwave Theory Tech.*, vol. 53, no. 3, pp. 1048-1055, Mar. 2005.
 - [145] I. H. Lin, M. DeVincentis, C. Caloz and T. Itoh, "Arbitrary dual-band components using composite right/left-handed transmission lines," *IEEE Trans. Microwave Theory Tech.*, vol. 52, no. 4, pp. 1142-1149, Apr. 2004.
 - [146] K. K. M. Cheng and F. L. Wong, "A novel approach to the design and implementation of dual-band compact planar 90 deg branch-line coupler," *IEEE Trans. Microwave Theory Tech.*, vol. 52, no. 11, pp. 2458-2463, Nov. 2004.
 - [147] R. F. Harrington, "Matrix methods for field problems," *Proc. IEEE*, vol. 55, no. 2, pp. 136-149, Feb. 1967.
 - [148] A. Gopinath and B. Easter, "Moment method of calculating discontinuity inductance of microstrip right-angle bends," *IEEE Trans. Microwave Theory Tech.*, vol. 22, no.10, pp. 880-883, Oct. 1974.
 - [149] P. Anders and F. Arndt, "Microstrip discontinuity capacitances and inductances for double steps, mitered bends with arbitrary angle, and

- asymmetric right-angle bends,” *IEEE Trans. Microwave Theory Tech.*, vol. 28, no. 11, pp. 1213-1217, Nov. 1980.
- [150] M. M. Ney, “Method of moments as applied to electromagnetic problems,” *IEEE Trans. Microwave Theory Tech.*, vol. 33, no. 10, pp. 972-980, Oct. 1985.
- [151] H. Y. Yang and N. G. Alexopoulos, “A dynamic model for microstrip – slotline transition and related structures,” *IEEE Trans. Microwave Theory Tech.*, vol. 36, no. 2, pp. 286-293, Feb. 1988.
- [152] R. N. Simons, N. I. Dib and L. P. B. Katehi, “Modeling of coplanar stripline discontinuities” *IEEE Trans. Microwave Theory Tech.*, vol. 44, no. 5, pp. 711-716, May 1996.