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Novel Methodologies For Miniaturized Filter Design And Realization

Ma KaiXue

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University

In fulfillment of the requirement for the degree of

Doctor of Philosophy

2007

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.

14/04/2007

Date

Ma Kaixue

Ma Kaixue

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Abstract

Tremendous growth in wireless communications has resulted in increasingly crowded frequency spectrum. As a result, a large number of passive components, such as high selectivity RF/microwave resonators and filters as well as baluns, are in demand in practical applications. Beside the performance, their compact size and low cost become very important of these passive components for the compact and portable communication systems. This thesis is therefore devoted to developing novel design methodologies for miniaturized resonators, filters and baluns with good performance. These designs are implemented and verified by using several fabrication technologies including print circuit board (PCB), thin-film, silicon micromachining and CMOS.

It is known that the miniaturization is the trend and designing compact passive devices is the challenging. Obviously, it would be impossible to face the challenging without novel design methodologies. Mainly, the electromagnetic (EM) characteristics determine the performance and size of the distributed transmission line filters. By investigating the EM characteristics of the circuits, different approaches, which form the basis of the thesis, are used to improve the performance and reduce the size of the filters and baluns.

Firstly, a novel general coupling method based on the investigation of the EM coupling paths is proposed. Instead of using the traditional approaches, that is, filter topology with non-cross-coupled resonators or filter topology with cross-coupled resonators, a new method is proposed to generate multiple zero-points. Using the proposed method, the inter-stage coupling between the adjacent resonators of the

filters can be well controlled. On the other hand, the filter size becomes much compact. To the best of author's knowledge, the canceling effects of the electric and magnetic couplings in two coupling paths have been explored and analyzed for the first time. Based on the above contributions, a new class of filter topology and microstrip filters with compact size and high performance are realized on PCB.

Secondly, the conductor-backed coplanar waveguide (CBCPW) slow-wave transmission lines and filters with gradual transition periodic loading are investigated extensively. A novel phase compensation method has been proposed for the design and analysis of the slow-wave filters which leads to the compact size and good stopband rejection. The proposed miniaturized filters have been implemented with the thin-film fabrication technology.

It is a challenge to realize filters on silicon substrates. The third contribution of the thesis is proposed a novel approach to design and implement the miniaturized filters on silicon wafers using self-developed SiDeox (Silicon deep etching and oxidation) processes. The proposed ground ring guarded patch resonators and filters with loading demonstrate high Q-factor and much compact size.

Finally, through investigating the electromagnetic characteristic of spirals on low resistivity silicon (resistivity of $\sim 15\Omega\cdot\text{cm}$), novel baluns have been proposed and implemented using commercial CMOS process. This work leads to the smallest balun in the world to cover the frequency range of 850MHz~2.5GHz.

Table of Contents

Statement of Originality	1
Acknowledgements.....	I
Abstract.....	II
Table of Contents.....	IV
List of Figures.....	i
List of Abbreviation	vi
CHAPTER 1	1
INTRODUCTION.....	1
1.1 Background and Motivation	1
1.2 Scope and Organization.....	5
References	7
CHAPTER 2	12
REVIEW OF DESIGNING TECHNIQUES AND BASIC CONCEPTS	12
2.1 Concepts of Filter	12
2.1.1 General definitions	12
2.1.2 Inverter for filter prototypes	14
2.1.3 Filter with non-cross-coupled resonators	18
2.1.4 Filter with cross-coupled resonators.....	21
2.2 Review of the Periodic Loading	23
2.3 Capacitive Loaded CBCPW Analysis.....	28
2.4 Conclusion.....	35
References	36
CHAPTER 3	41
COMPACT FILTER DESIGN ON PCB.....	41
3.1 Open-Ground Structure	44
3.2 Coupling Characteristics of the OG Structure	47
3.3 Model-Based Design of Tap-Feed OG Filter	53
3.3.1 Scalable model	55

3.3.2	Filter design	57
3.3.3	Results and discussion.....	60
3.4	U-shape Filter and Diplexer	63
3.4.1	Filter characterizing.....	64
3.4.2	Analytical modeling	67
3.4.3	Filter and diplexer	72
3.5	Spurious Suppression.....	78
3.5.1	Characteristics of resonator	80
3.5.2	Filter design	82
3.5.3	Experiments and results	84
3.6	A Compact Size Coupling Controllable SEMCP Filter	86
3.6.1	Analysis of OG filters with SEMCP	88
3.6.2	Second-order OG filter realization	95
3.6.3	General higher order SEMCP OG filter	101
3.7	Conclusion.....	107
	References	108
CHAPTER 4.....		114
PERIODIC LOADING LINES AND FILTERS USING THIN-FILM TECHNOLOGY.....		114
4.1	Parameter extracting of the Unit Cells.....	116
4.1.1	Parameter extracting of the transmission cells.....	117
4.1.2	Parameter extracting of the coupling cells.....	120
4.2	Phase Compensation of CBCPW Lines	126
4.2.1	Discussion on leakage modes	129
4.2.2	Phase compensation method	131
4.3	Loaded and Unloaded CBCPW Filters	135
4.3.1	Filter design	135
4.3.2	Discussions on power leakage of the CBCPW filters	138
4.3.3	Size reduction factor of the CBCPW filters	141
4.4	Miniaturized Filters Using Zigzag Lateral Grounds	143
4.4.1	Configuration	143

4.4.2	Discussion on lateral grounding effects	147
4.5	Conclusion.....	148
	References	149
CHAPTER 5		153
FILTER DESIGN USING SILICON-BASED MICROMACHINING TECHNOLOGY.....		153
5.1	CBCPW Feed CBCPS Ring Resonator	154
5.1.1	General ring circuit.....	154
5.1.2	Ring resonator description	155
5.2	A Planar Miniaturized High-Pass Filter	161
5.2.1	Circuit description	161
5.2.2	Filter results	163
5.3	A Ground Guarded Slit Patch Resonator	164
5.3.1	General patch circuit	164
5.3.2	Ground guarded patch resonator with slits	165
5.4	Ground-Guarded Patch Resonator and Filter.....	169
5.4.1	Guarded patch resonators	170
5.4.2	Guarded patch filter with ground shunt	172
5.5	Conclusion.....	176
	References	176
CHAPTER 6		181
RESONATOR AND BALUN DESIGNS USING THE STANDARD CMOS TECHNOLOGY		181
6.1	Ka Band CMOS DMML Ring Resonator	181
6.1.1	Dual-meander line ring.....	183
6.1.2	Realization and measurement	186
6.2	Design of Miniaturized Stacked Baluns on CMOS	189
6.2.1	Balun construction and design	190
6.2.2	Single-ended balun realization	192
6.2.3	Single-ended drive 800 MHz~2.5 GHz balun.....	194
6.2.4	Single-ended drive 1.5GHz~4GHz balun	196

6.2.5	Parasitic effects analysis	198
6.2.6	Balun loss reduction	202
6.3	Conclusion	205
	References	206
CHAPTER 7		210
SUMMARY AND RECOMMENDATIONS		210
7.1	Summary	210
7.2	Recommendations	213
Author's Information		215

List of Figures

Figure 1.1 RF passive circuit design linkages	4
Figure 2.1 Low-pass filter prototypes.....	15
Figure 2.2 Definition of impedance and admittance inverter	15
Figure 2.3 The prototype elements and their values for filters	17
Figure 2.4 Equations for two port half-wavelength resonator filter	19
Figure 2.5 Equations for the two port quarter-wavelength resonator filter.....	20
Figure 2.6 The general coupling topology of the filter.....	23
Figure 2.7 Thin diaphragm shunt capacitive loading of a CBCPW line.....	29
Figure 2.8 The equivalent circuits for the unit cell and cascade cells.....	30
Figure 3.1 Proposed OG configuration.....	44
Figure 3.2 The voltage distribution on the half-wavelength line.....	45
Figure 3.3 Equivalent circuits of odd- and even- Modes.....	47
Figure 3.4 Coupling coefficient and resonance frequency of OG structure	50
Figure 3.5 OG structures with four kinds of perturbations.....	51
Figure 3.6 Equivalent circuit of perfected conducting via.....	52
Figure 3.7 Comparison between the coupling coefficients of OLV and BOLV.....	52
Figure 3.8 Comparison of the coupling coefficients of ILV and PV	53
Figure 3.9 Configuration and the model of the proposed filter	54
Figure 3.10 Comparison of model-based results with and without open end effects	60
Figure 3.11 Comparison of the results with and without open-end effects.....	60
Figure 3.12 Comparison of the model-based results and experimental results.....	61
Figure 3.13 Photograph of the fabricated BPF	62
Figure 3.14 Comparison of the theory results with and without electric coupling	62
Figure 3.15 Responses of the different type second-order BPFs.....	63
Figure 3.16 a) The compact OG filter b) Filter topology.....	64
Figure 3.17 The external Q -factor of a resonator with respect to length L_l	65
Figure 3.18 OG filters with 6% bandwidth	66
Figure 3.19 OG filter with 18% bandwidth.....	67

Figure 3.20 The equivalent circuit model.....	67
Figure 3.21 Response of equivalent circuit of the OG Filter	69
Figure 3.22 The ZP positions in the spectrum versus coupling capacitance C_m	69
Figure 3.23 The relationship among the transmission Zps, C_m and Δ	69
Figure 3.24 The EM results and the lumped model results.	72
Figure 3.25 Comparison between EM simulated results from IE3D and the measurement	73
Figure 3.26 Comparison between the EM simulated results from HFSS and the measurement	74
Figure 3.27 Photograph of the filter	74
Figure 3.28 The configuration of the diplexer.....	75
Figure 3.29 Simulation results of the diplexer.....	77
Figure 3.30 Measured results of the diplexer	77
Figure 3.31 Photographs of the diplexer.....	78
Figure 3.32 Quarter-wavelength resonators with different feeder types.....	79
Figure 3.33 Configuration for characterizing the resonator with “T” feeder.....	80
Figure 3.34 Comparison of the transmission under different feeder dimensions	81
Figure 3.35 Configuration of the filter with spurious suppression	83
Figure 3.36 Comparison of the theory and experiment results of the proposed filter.....	84
Figure 3.37 Passband performance of proposed filter C) of Figure 3.36.....	84
Figure 3.38 Stopband performance of proposed filter C) of Figure 3.36	85
Figure 3.39 The enlarged photograph of the filter with spurious suppression.....	86
Figure 3.40 Proposed second-order filter configuration and the topology	88
Figure 3.41 The equivalent even-mode configuration.....	89
Figure 3.42 The equivalent odd-mode configuration	90
Figure 3.43 The spectrum responses of M dominant SEMCP filters.....	93
Figure 3.44 The spectrum responses of E dominant SEMCP filters.....	94
Figure 3.45 The configurations of the second-order SEMCP filters.....	95
Figure 3.46 The frequency responses of the second-order E dominant SEMCP filters.....	97
Figure 3.47 The frequency responses of the second-order M dominant SEMCP filters.....	98
Figure 3.48 Simulated and measured results of second-order M dominant filter	99
Figure 3.49 The photograph of the fabricated M dominant SEMCP filter.....	99

Figure 3.50 Simulated and measured results of the second-order E dominant filter	100
Figure 3.51 Photograph of the E dominant second-order SEMCP filter	100
Figure 3.52 The topology of the high-order SEMCP filter	101
Figure 3.53 The configuration and topology of the fourth-order SEMCP filter	102
Figure 3.54 The configuration and topology of the fourth-order SEMCP filter	104
Figure 3.55 Simulated and measured results of the fourth-order E dominant filter.....	105
Figure 3.56 Measured results of the fourth-order SEMCP filter and hairpin filter	106
Figure 3.57 The photograph of the fabricated fourth-order SEMCP and hairpin filters	107
Figure 3.58 Group delays of the measured fourth-order SEMCP and hairpin filters	107
Figure 4.1 Structure of conductor-backed coplanar waveguide (CBCPW)	114
Figure 4.2 Transmission cells and their lumped models	117
Figure 4.3 Phase characteristics of the three kinds of cells	118
Figure 4.4 Frequency-dependent element values of π -network models of T-cell	119
Figure 4.5 Coupling cells and the lumped models:	121
Figure 4.6 Frequency-dependent lumped element values of CC cell models.....	122
Figure 4.7 Frequency-dependent lumped element values of IC cell models	123
Figure 4.8 Normalized K -inverter value and equivalent electric length of IC cell	124
Figure 4.9 Normalized J -inverter value and equivalent electric length of CC cell.....	125
Figure 4.10 Top metal patterns of the 9-cells CBCPW transmission lines	126
Figure 4.11 Measured transmission characteristics of periodic cell lines.....	128
Figure 4.12 Measured reflection characteristics of periodic cell lines	128
Figure 4.13 Measured phase delay of periodic cell lines.....	128
Figure 4.14 Substrate reflections and radiations of PP mode in finite CBCPW	129
Figure 4.15 Relationship of the cell number with possible resonant frequency	130
Figure 4.16 Measured results of the lines in Figure 4.10 with and without ground via.....	131
Figure 4.17 Cell cascaded transmission line with p cells and the errors.....	132
Figure 4.18 The first-order and the second-order errors of the cascaded cells	133
Figure 4.19 Comparisons of the phase delay error between direct cascade and full- wave EM results and between cascade with compensation and full-wave EM results.....	134
Figure 4.20 Phase delay errors between the calculations and measurements	135

Figure 4.21 Configurations of second-order CBCPW filters.....	136
Figure 4.22 Cell topology of second-order CBCPW filter in Figure 4.21 b).....	136
Figure 4.23 Measured results of CBCPW filters with and without loading	138
Figure 4.24 Measured results of IT loaded and the unloaded BPFs	138
Figure 4.25 Measured results of the CT loaded filters with and without via.....	139
Figure 4.26 Measured results of the IT loaded filters with and without via	139
Figure 4.27 Measured loss factors of the CT loaded filters with and without via	141
Figure 4.28 Measured loss factors of the IT loaded filters with and without via.....	141
Figure 4.29 Enlarged photographs of fabricated lines and filters	142
Figure 4.30 Perspective view of zigzag lateral ground CBCPW filter	143
Figure 4.31 Top metal of the CBCPW filters	144
Figure 4.32 Simulation results of the filters in curves a) and b) of Figure 4.31	145
Figure 4.33 Simulated and measured results of the filter in Figure 4.31 b).....	146
Figure 4.34 Results of the filter with 4-pair metal via holes in narrow band range.....	147
Figure 4.35 Comparison of filters with different shielding types	148
Figure 4.36 Enlarged photographs of filters	148
Figure 5.1 The top view of the CBCPW feed CBCPS ring resonator	155
Figure 5.2 Cross section of three-layer dielectric transmission lines	156
Figure 5.3 Comparison of the MS and CBCPS rings in narrow frequency range	157
Figure 5.4 Comparison of the MS ring and CBCPS ring in wide frequency range.....	158
Figure 5.5 Current directions of the CBCPS ring resonator	159
Figure 5.6 Measured spectrum response of Via-CBCPW line.....	159
Figure 5.7 Measured spectrum response of CBCPS ring resonator.....	160
Figure 5.8 Proposed HCTL high pass filter	162
Figure 5.9 Simulation result of the HCTL high pass filter	164
Figure 5.10 Measured results of the HCTL high pass filter.....	164
Figure 5.11 Configuration of the ground guarded slit patch resonator/filter	166
Figure 5.12 Comparison of the response of three patch resonator/filter.....	167
Figure 5.13 Comparison of the simulation and measurement	168
Figure 5.14 Realization of the filter.....	169

Figure 5.15 Perspective view of ring guarded shunt patch filter	170
Figure 5.16 Ground ring guarded patch resonators	171
Figure 5.17 The frequency response of the resonators in Figure 5.16	172
Figure 5.18 Comparison of Figure 5.16 a) and Figure 5.16 b) type filters	173
Figure 5.19 The shielding effects on ring guarded shunt patch filter	174
Figure 5.20 Comparison of the theory and experiment results of the proposed filter.....	175
Figure 5.21 Comparison of the theory and experiment loss factor of the filter	175
Figure 5.22 Enlarged photograph of the proposed filter in Figure 5.15	175
Figure 6.1 CMOS dual-meander lines ring resonator.....	182
Figure 6.2 Top view of the meander line	183
Figure 6.3 The transmission characteristics of meander line.....	184
Figure 6.5 The DMML ring resonator under test	187
Figure 6.6 Measured spectrum response of DMML ring.....	188
Figure 6.7 the construction of the MSSB	191
Figure 6.8 Simulation results of MSSB under differential drive	192
Figure 6.9 Multilayer balun configuration (not to scale).....	193
Figure 6.10 The balun wafer in the chamber measurement system.....	194
Figure 6.11 Measured results of single-ended drive balun A.....	195
Figure 6.12 Measured and model-based results of single-ended drive balun A	196
Figure 6.13 Die photograph of single-ended drive balun A with guard ring	196
Figure 6.14 Measured results of single-ended drive balun B	197
Figure 6.15 Measured and model-based results of single-ended drive balun B	197
Figure 6.16 Single-ended drive Multilayer balun model.....	199
Figure 6.17 Comparison of the symmetrical and asymmetrical baluns.....	202
Figure 6.18 Schematic of single-ended drive balun with tuning capacitors	203
Figure 6.19 Balun characteristics under two different tuning conditions	204
Figure 6.20 Comparison of the asymmetrical balun with and without tuning.....	205

List of Abbreviation

GSM	global system of mobile communication
PCS	personal communication system
W-LAN	wireless local area network
SOCs	system-on-chips
MMIC	monolithic microwave integrated circuits
RF-MEMS	RF microelectromechanic systems
LTCC	low-temperature co-fired ceramics
PCB	print circuit board
CMOS	complementary-symmetry metal–oxide–semiconductor
EM	electromagnetic
CPW	coplanar waveguide
CBCPW	conductor-backed coplanar waveguide
CBCPS	conductor-backed coplanar strip
SiDeox	silicon deep etching and oxidation
ZP	zero point
SEMCP	separate electric and magnetic coupling paths
DECT	digital enhanced cordless telecommunication
MIC	microwave integrated circuit
MS	microstrip
BPF	band pass filter
OG	open-ground
IG	imperfect ground
GR	ground reference
OPD	open-ended
BW	bandwidth
FBW	fractional bandwidth
HRB	high rejection band
LRB	low rejection band

SIR	stepped-impedance resonator
T-cell	transmission cell
C-cell	coupled cell
CT	capacitive transmission
IT	inductive transmission
CC	capacitive coupling
IC	inductive coupling
PP	parallel plate
CPS	coplanar strips
HCTL	via-CBCPW cascaded with three coupling MS lines
DMML	dual microstrip meander line
I/O	input/output
DUT	device under test
MIM	metal-isolator-metal
MSSB	multilayer symmetrical stacked balun
CSM	chartered semiconductor manufacturing ltd.
FSS	frequency selective surface
EBG	electromagnetic band gap
PBG	photonic band gap
TLM	transmission line matrix
SAW	surface acoustic wave
PIFA	planar inverted-F antenna
CQ	cascade quadruplet
CT	cascade trisection

CHAPTER 1

INTRODUCTION

1.1 Background and Motivation

The increasing worldwide demands for a variety of information and communication services greatly stimulate the development of contemporary communication systems. The wireless communication, undergoing a revolution during the twentieth century especially since the first commercial use of the satellite communication in 1965 [1]-[2], becomes one of the most promising areas of growth in the 21st century [3]. The demands for ubiquitous communications have led to the development of various mobile/wireless systems, including GSM (global system of mobile communication), PCS (personal communication system), W-LAN (wireless local area network), Bluetooth, etc., which are evolving towards the convergence of disparate networks and standards including all IP (internet protocol) network, interoperable networks and even satellite networks of four orbit types [4]. The technology advancement and new frequency allocation for the new communication systems are creating the numerous business and technique opportunities. Generally speaking, the mobile/wireless communications are in many ways penetrating and changing the daily lives of the people all over the world.

However, tremendous growth in wireless communications has greatly crowded the frequency spectrum, which may transfer into a higher likelihood of user interference with one another. To prevent interferences among the channels, ensure high quality

communications and prevent the potential frequency interferences due to the nonlinear characteristics in the systems, high selectivity and high rejection are required not only between the adjacent channels but also in the same channel [5]-[6]. In addition, in order to be accepted by the market, the communication systems with fast services, low cost and compact size are necessary. As a result, the components for the systems, especially for those used for frequency selection, become extremely difficult to design with ever more stringent requirements [7].

The RF/microwave filters play vital roles in the frequency selection, rejecting unused or interference frequencies in the systems and transmitting certain frequency ranges without attenuation, in almost all of the wireless communication systems. The performance of the RF/microwave filters, which mainly determines the quality of the RF signal, indirectly affects the design simplicity, reliability and cost of base-band parts, front-end parts or the entire systems. On the other hand, a large portion of the area in the portable communication systems is occupied by the RF/microwave passive circuits such as filters, baluns, etc. Thus, a compact size of these passive circuits becomes another key factor considered by the system designer. This thesis is therefore devoted to investigating suitable techniques to design and implement RF/microwave passive filters with a compact size and high performance.

In order to obtain good performance, differential architectures are popular in most wireless communication transceivers [8]-[10]. Therefore, a balun, providing balanced differential outputs from the single-ended input signal, is needed. There are numerous publications on various baluns in the literature [11]-[17]. Distributed

passive baluns, like Marchand baluns, which consist of sections of half wavelength or quarter wavelength transmission lines [8]-[9] or coupled lines [10] at the center operating frequency, occupy large areas. Thus, these transmission line sections are replaced by meandering lines or inter-wound spiral transformers to achieve more compact size [11]-[12]. Lumped element baluns, employing low pass and high pass filter structures [14]-[15] or transformers, have also been demonstrated. Unfortunately, these structures require a lot of lumped components and the higher frequency is limited by the parasitic effects of lumped elements. The balun structures employing both lumped and distributed elements have been reported in [16]-[17], where the lumped elements are used to reduce the length of the section line to much lesser than a quarter wavelength [18]. But, all those reported single-ended drive baluns are not realized on standard CMOS technology. In [19], the single-ended drive baluns are realized on high resistivity silicon ($>4\text{K}\Omega\cdot\text{cm}$). However, architectures proposed by the authors of [17] failed to be realized in the high loss substrate (low resistivity $\sim 15\Omega\cdot\text{cm}$). Moreover, these reported balun configurations are still too large to be economically realizable by CMOS process. The trend of Tx/Rx front-ends in wireless communication is to use less off-chip components, particularly, to be so-called system-on-chips (SOCs). Therefore, to realize a miniaturized balun on CMOS will be one of the key innovations in SOC's implementation. The second objective of this dissertation is to develop a compact size balun directly based on commercial CMOS technology.

Figure 1.1 RF passive circuit design linkages

The passive circuits based on traditional printed circuits board (PCB) are still drawing much attention due to their low cost and new emerging design configurations and technologies [7], [19]-[20]. The recent advancement of novel materials and fabrication technologies including monolithic microwave integrated circuits (MMIC) [21], RF microelectromechanic systems (RF-MEMS) [22]-[24] and micromachining [25]-[26], high-temperature superconductor [20], [27], low-temperature co-fired ceramics (LTCC) [28]-[29] and photonic bandgap materials/structures [20], [30]-[31] etc., as illustrated in Figure 1.1, has stimulated the rapid development of the new design or realization of the RF/Microwave passive circuits such as filters and baluns. Actually, all of these technologies have their advantages and disadvantages in realizations of the filter circuits. For practical utilization, most of the systems are developed under the orientation of performance per unit cost. The investigation of the circuits on different fabrication technologies is not only helpful to understand such

available technologies but also give more choices for system engineers to meet better design specifications. Therefore, the third and final objective is devoted to investigating passive circuit design such as filter and balun with several different technologies including PCB, micromachining on silicon, thin-film and CMOS technologies.

1.2 Scope and Organization

This thesis focuses on the analysis of EM characteristics, design and implementation of passive filters and baluns using several available fabrication technologies. As shown in Figure 1.1, the electromagnetic (EM) theory, circuit theory and material characteristics are essential to investigate and realize these passive devices. The methods discussed in this thesis include full-wave EM simulations, transmission line analysis, filter principle, odd- and even- mode analyses, EM based parameter extraction and the microwave circuit modeling. The microwave structures in this thesis cover microstrip lines, conductor-backed coplanar waveguide (CBCPW), conductor-backed coplanar strip (CBCPS) and different microwave resonator circuits including ring resonator, patch resonator, half-wavelength resonator, quarter-wavelength resonator, etc. The designs, covering the novel topology design and configuration design, deal with the structures from the single layer substrate with double metal to multiple substrate layers with multiple metal layers (up to six substrate layers with six metal layers). Moreover, different fabrication technologies including PCB, thin-film ceramics, silicon micromachining and CMOS are used to

implement these devices. Measurements including on-board measurement as well as probe-tip measurement are performed to verify the design and analysis.

The organization of the thesis is as follows:

The fundamental principles and concepts for the periodic loading and filter design are briefly introduced in Chapter 2.

In Chapter 3, several novel resonator and filter configurations and topologies are proposed and implemented using conventional PCB technology. The methodologies for filter size reduction, stopband extension and multiple zero points (ZPs) generation are introduced and demonstrated theoretically and experimentally. A novel SEMCP (separate electric and magnetic coupling paths) filter and its general topology are also introduced in this chapter.

The investigation and realization of novel periodic structures and filters using the thin film technology are presented in Chapter 4. The lumped circuit model parameters extracted from the EM results of the cell and compensation method for directly cells cascading are demonstrated and used for the filter designs. The “via ground” approach and the “finite ground” approach for free power leakage corresponding to CBCPW slow-wave lines and filters are also discussed in this chapter. Eventually, a novel CBCPW bandpass filter (BPF) using zigzag lateral ground, coupling enhancement and leakage suppression techniques, are implemented.

In Chapter 5, several resonators and filters are investigated and implemented on a silicon wafer by using so-called SiDeox (silicon deep etching and oxidation) and through hole plating technology. The transmission and shielding characteristics of the

ring guarded resonator with ground shunt are also investigated.

In Chapter 6, a resonator, ultra-wide band transformer and single-ended drive multilayer stacked baluns are investigated and implemented on 0.18 μ m CMOS technology. The equivalent-circuit model of the asymmetrical configuration and the resonant tuning method are proposed and used to analyze and improve the balance and the loss characteristics of the baluns.

Finally, the thesis is summarized and concluded in Chapter 7. Discussions and suggestions of future work are also discussed.

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CHAPTER 2

REVIEW OF DESIGNING TECHNIQUES AND BASIC CONCEPTS

This chapter includes two parts, that is, the concepts of filter and the technique of periodic loading. Although the physical realization of filters at RF/Microwave frequencies may vary, the circuit network topology is common to all. Therefore, the technique content of the thesis begins with section 2.1.1, which describes the network parameters and equations for the filter. On the other hand, the inverters are widely adopted to transform the low pass prototype filter to bandpass filter. Therefore the low pass prototype filter and the inverter, which are used in the following chapters, are introduced in section 2.1.2. The bandpass filter with resonators can be classified as two groups, that is, filter with non-cross-coupled resonators and filter with cross-coupled resonators. These two types of filters are introduced in section 2.1.3 and section 2.1.4 respectively. Since the periodic loading is the basis of Chapter 4, the literature review and theory analysis of the periodic loading are presented in section 2.2 and section 2.3 respectively.

2.1 Concepts of Filter

2.1.1 General definitions

In general, a filter is a two port network. The transfer function of the two-port filter network is a mathematical description of network response characteristics,

namely, a mathematical expression of S_{21} . On many occasions, an amplitude-squared transfer function for a lossless passive filter network is defined as [1]

$$|S_{21}(\omega)|^2 = \frac{1}{1 + \Delta^2 F_n^2(\omega)} \quad (2.1)$$

where Δ is the ripple constant related to the passband return loss L_R . It is defined as

$$\Delta = [10^{L_R/10} - 1]^{-1/2}$$

$F_n^2(\omega)$ represents a filtering or characteristic function. ω is a frequency variable.

For filter characterization, two parameters need to be defined [1]

$$L_A = -20 \log |S_{21}| \quad \text{dB} \quad (2.2a)$$

and

$$L_R = -20 \log |S_{11}| \quad \text{dB} \quad (2.2b)$$

where L_A denotes the insertion loss between ports 1 and 2 and L_R represents the return loss at port 1.

For linear, time-invariant networks, the transfer function may be defined as a rational function, that is

$$S_{21} = \frac{N(q)}{D(q)} \quad (2.3)$$

where $N(q)$ and $D(q)$ are polynomials with complex frequency variable $q = \sigma + j\omega$.

For a lossless passive network, the neper frequency $\sigma = 0$, hence $q = j\omega$. To find a realizable rational transfer function that produces response characteristics approximating the required response is the so-called approximation problem, and in many cases, the rational transfer function of (2.3) can be constructed from the amplitude-squared transfer function of (2.1).

For a given rational transfer function of (2.3), the insertion loss response of the filter, following the conventional definition in (2.2), can be computed by [1]

$$L_A(\omega) = 10 \log \frac{1}{|S_{21}(j\omega)|^2} \quad \text{dB} \quad (2.4)$$

Since $|S_{11}|^2 + |S_{21}|^2 = 1$ for a lossless, passive two-port network, the return loss response of the filter can be found using (2.2)

$$L_R(\omega) = 10 \log \left[1 - |S_{21}(j\omega)|^2 \right] \quad \text{dB} \quad (2.5)$$

If a rational transfer function is available, the phase response of the filter can be found as

$$\phi_{21} = \arg S_{21}(j\omega) \quad (2.6)$$

Then the group delay response of this network can be calculated by

$$\tau_d(\omega) = \frac{d\phi_{21}(\omega)}{-d\omega} \quad \text{seconds} \quad (2.7)$$

where $\phi_{21}(\omega)$ is in radians and ω is in radians per second.

2.1.2 Inverter for filter prototypes

The element values of low-pass prototype filters [2] are defined in Figure 2.1.

The related elements in prototype filter are

$$\begin{aligned} g_k \big|_{k=1 \text{ to } n} &= \begin{cases} \text{the inductance of a series coil} \\ \text{or the capacitance of shunt capacitor} \end{cases} \\ g_0 &= \begin{cases} \text{the generator resistance } R_0' & \text{if } g_1 = C_1' \\ \text{the generator conductance } G_0' & \text{if } g_1 = L_1' \end{cases} \\ g_{n+1} &= \begin{cases} \text{the load resistance } R_{n+1}' & \text{if } g_n = C_n' \\ \text{the load conductance } G_{n+1}' & \text{if } g_n = L_n' \end{cases} \end{aligned}$$

One possible form of a prototype filter is shown in Figure 2.1 a), while its' dual

is shown at Figure 2.1 b). Since both give identical responses, either form may be used. Since the networks are reciprocal, either the resistor on the left or the one on the right may be defined as the generator internal impedance.

Figure 2.1 Low-pass filter prototypes

a) prototype circuit b) the dual of a)

Figure 2.2 Definition of impedance and admittance inverter

In deriving design equations for certain types of band pass and band stop filters, it is desirable to convert the low pass filter prototypes, which use both inductances

and capacitances, to equivalent forms which use only inductances or only capacitances. This can be done with the aid of the idealized inverters which are symbolized in Figure 2.2.

An ideal impedance inverter [2] operates like a quarter-wavelength line with the characteristic impedance of $K_{k,k+1}$ at all frequencies. Therefore, if it is terminated with an impedance of Z_b on one end, the impedance Z_a seen from the other end is

$$Z_a = \frac{K_{k,k+1}^2}{Z_b} \quad (2.8)$$

An ideal admittance inverter as defined herein is the admittance representation of the same thing, i.e., it operates like a quarter-wavelength line with characteristic admittance of $J_{k,k+1}$ at all frequencies. Thus, if an admittance Y_b is attached at one end, the admittance Y_a seen looking in other end is

$$Y_a = \frac{J_{k,k+1}^2}{Y_b} \quad (2.9)$$

As indicated in Figure 2.2, an inverter may have an image phase shift of either ± 90 degrees or odd multiples thereof.

Because of the inverting transformation indicated by equations (2.8) and (2.9), a series inductance with an inverter on each side looks like a shunt capacitance from exterior terminals. Likewise, a shunt capacitance with an inverter on both sides looks like a series inductance from its external terminals. Making use of this property, the prototype circuits in Figure 2.2 can be converted into either of the equivalent forms given in Figure 2.3 which have identical transmission characteristics to those low pass prototypes. Observed from equations (2.8) and (2.9), inverters have the ability to shift

impedance or admittance levels depending on the choices of K and J parameters. For this reason, the size of R_A , R_B and the inductances L_{0k} ($k=1,2,3,\dots,n$) in Figure 2.3 may be chosen arbitrarily. The response will be identical to that of the original prototype as shown in Figure 2.1 provided that the inverter parameters $K_{k,k+1}$ are specified as indicated by the equations in Figure 2.3 a). The same condition holds for the circuit in Figure 2.3 b) only on the dual basis. Note that the g_k values referred to the equations in Figure 2.3, the element values of the prototype are defined in Figure 2.1.

Figure 2.3 The prototype elements and their values for filters

A method to derive the equations for K and J inverters will now be briefly explained. A fundamental method of looking at the relation between the prototype circuits in Figures 2.1 a), b) and the corresponding circuit in Figure 2.3 a) makes use of the concept of duality. A given circuit as seen through an impedance inverter looks like the dual of that given circuit. Thus, the impedances seen from inductor L_{01} in

Figure 2.3 a) are the same as those seen from inductance L_1' in Figure 2.1 b), except for an impedance scaling factor. The impedance seen from inductor L_{02} in Figure 2.3 a) is identical to those seen from inductance L_2' in Figure 2.1 a), except for possible impedance scaling change. In this manner the impedance at any point of the circuit in Figure 2.3 a) may be quantitatively related to the corresponding impedances in circuits in Figure 2.1 a), b) [2].

2.1.3 Filter with non-cross-coupled resonators

Filter with non-cross-coupled resonators is a filter with inter-stage coupling only between the adjacent resonators in space. The end-coupled filter, which includes the half-wavelength filter and the quarter-wavelength filter, is a kind of the filter with non-cross-coupled resonators. Two kinds of half-wavelength resonators are widely used in end-coupled filter with half-wavelength resonators. One is that both terminals of the resonator are shorted, and the other one is that both terminals are opened. The gap capacitance coupled structures and shunt inductance coupled structures, as shown in Figure 2.4 a) and b) respectively, are used to build two different kinds of half-wavelength band pass filters (BPF). These filters consist of transmission-line resonators which are approximately a half-wavelength long at the center frequency f_0 and which have series capacitance coupling or shunt inductance coupling between the resonators. The design equations [2] are given in Figure 2.4

Figure 2.4 Equations for two port half-wavelength resonator filter

where $g_0, g_1, \dots, g_n$ are as defined in Section 2.1.2, ω_1' is the passband cutoff frequency and w is the fractional bandwidth. w and ω' are used to approximately map the low-pass prototype filter to the corresponding BPF response. They are defined as

$$w = 2 \left(\frac{\omega_2 - \omega_1}{\omega_2 + \omega_1} \right) \quad (2.10)$$

$$\omega' = \frac{2\omega_1'}{w} \left(\frac{\omega - \omega_0}{\omega} \right) \quad (2.11)$$

where ω_1 and ω_2 are low and upper band-edge frequency of the BPF respectively and the centre frequency is given as

$$\omega_0 = \frac{2\omega_2\omega_1}{\omega_2 + \omega_1} \quad (2.12)$$

Impedance inverters corresponding to coupled shunts are used in the filter with coupled shunts in Figure 2.4 a) where $Z_0 = 1/Y_0$ is the characteristic impedance of the line between the inverters. In the filter with coupled gaps shown in Figure 2.4 b), the admittance inverters corresponding to coupled gap are used for the filter design. $K_{j,j+1}$,

$J_{j,j+1}$ and $\phi_{j,j+1}$ of the coupling structure, which are corresponding to X , B and ϕ in different sections ($j=1,2,3,\dots,n$) respectively, can be computed from the equations in Figure 2.4. Then

$$\theta_j = \pi + \frac{1}{2}[\phi_{j-1,j} + \phi_{j,j+1}] \quad \text{radians} \quad (2.13)$$

where $\phi_{j-1,j}$ and $\phi_{j,j+1}$ are obtained from the transformation from the admittance of X to the K inverter or from the admittance of B to the J inverter.

If the quarter-wavelength resonators are coupled alternately by K - and J -inverters, the BPF in Figure 2.5 can be constructed. Although other types of construction and other types of K - and J -inverters may also be used, Figure 2.5 gives the general design data for TEM-mode type of filter. In this structure, impedance inverter alternates with the admittance inverters. $Z_0 = 1/Y_0$ is the characteristic impedance of the line between the inverters. $K_{j,j+1}$, $J_{j,j+1}$ and $\phi_{j,j+1}$ of the coupling structure can be computed from the equations in Figure 2.5 and the effective electrical length can be calculated by [2]

$$\theta_j = \frac{\pi}{2} + \frac{1}{2}[\phi_{j-1,j} + \phi_{j,j+1}] \quad \text{radians} \quad (2.14)$$

Figure 2.5 Equations for the two port quarter-wavelength resonator filter

2.1.4 Filter with cross-coupled resonators

Filter with increasingly stringent requirements can often be realized only by using cross-coupled resonators to generate finite transmission zeros. A general theory of cross-coupled resonator bandpass filters was developed in 1970s by Atia and Williams [3]-[7]. Lower order filters, up to fourth-order, were solved analytically by Kurzrok [6], [8] and Williams [7]. Cameron also gave a scheme to determine the filtering function with arbitrarily placed transmission zeros [8]-[10]. Once the system function is obtained, the synthesis of the filter proceeds with extracting element values [9] to obtain a coupling matrix. Other excellent techniques were also presented by many researchers, most notably by the groups around Rhodes [11]-[16]. The literature on this subject is too extensive to be listed here. A good review can be found in special issue [17]. The theory by Atia and Williams leads to a coupling matrix which reproduces the system function to be synthesized but often includes unwanted or unrealizable coupling elements. Repeated similarity transformations are then used to cancel the unwanted couplings [4], [10]. Unfortunately, the process does not always converge [18]. The same approach was also used in a recent publication by Cameron [19] to reduce a potentially full coupling matrix to a folded form. Recently, optimization was also used in synthesizing this type of microwave structure [8], [17]-[18].

A typical structure considered in [7] is shown in Figure 2.6. It consists of N coupled lossless resonators. The resonant frequency is assumed to be unity. The frequency-independent coupling coefficient between resonators i and j is denoted by

$M_{ij}=M_{ji}$. Possible shifts in the resonant frequencies of the resonators are included in the diagonal elements of the coupling matrix $[M]$. A voltage source of magnitude equal to unity and internal resistance R_1 excites the structure at resonator 1. The load at the output is a resistor R_2 connected to resonator N . The loop currents in the different resonators are grouped in the vector $[I]$. Following the analysis in [4], the loop currents are governed by the following matrix equation:

$$[\lambda - jR + M][I] = [A][I] = -j[e] \quad j^2 = -1, \quad \lambda = \omega - \frac{1}{\omega} \quad (2.15)$$

where ω is the normalized angular frequency. R is a matrix whose only non-zero entries are $R_{11}=R_1$, and $R_{NN}=R_2$. M is a symmetric square coupling matrix (with zero diagonal elements when only synchronously tuned resonators are considered). The excitation vector $[e]$ is given by $[1,0,0,\dots,0]^t$, where t is the transposition operator. The discussion on the limitations of this model to narrow-band filters is well presented in [4] and is not repeated here.

The response functions we are interested in are the insertion and return loss of the filter as a function of frequency. Using equation (2.15), we get the loop currents as

$$[I] = -j[A^{-1}] [e] \quad (2.16)$$

From a simple analysis of the circuit, we get the following expressions for the transmission and reflection coefficients:

$$S_{21} = 2\sqrt{R_1 R_2} I_N = -2j\sqrt{R_1 R_2} [A^{-1}]_{N,1} \quad (2.17)$$

and

$$S_{11} = 1 - 2R_1 I_1 = 1 + 2jR_1 [A^{-1}]_{1,1} \quad (2.18)$$

The coupling matrix and the termination resistances are assumed to be determined

using the theories in [4]-[8].

Figure 2.6 The general coupling topology of the filter

From Figure 2.6, it can be seen that the filter with non-cross-coupled is a special case of the filter with cross-coupled resonators when the coupling coefficients between non-adjacent resonators are zero. We will demonstrate a novel coupling method based on the investigation of the EM coupling paths in Chapter 3. Instead of using the filter topology with cross-coupled resonators introduced above, multiple zero-points can be obtained by applying the proposed method in Chapter 3.

2.2 Review of the Periodic Loading

The periodic and guiding structures [20]-[33] play an important role in the operation of the components like antennas, electron tubes, low-noise amplifiers and filters, etc. Periodic structures have drawn much attention on the waveguide, microwave integrated circuit (MIC), MMIC and RF-MEMS applications due to following four important characteristics:

1) Periodic guiding structures can be considered as one kind of reactively periodic loading which leads to the distinctive passband and stopband characteristics in

electromagnetic propagation.

2) Due to the reactively periodic loading, the transmission of electromagnetic wave supports the so-called slow-wave propagation which can be used to reduce the size of the components.

3) Periodic structures can be used to form the electromagnetic band gap (EBG) or photonic band gap (PBG) structures. Some passive components such as leaky wave antenna, frequency selective surface (FSS) and filters use this characteristic to improve the performance of the circuits.

4) Periodic structures are easier to be analyzed and modeled as compared to designs based on non-periodic characteristics.

In order to reveal the importance of the periodic structures in communication and to highlight the above mentioned advantages, several typical application cases, operating principles and examples are briefly introduced in the following:

a) Periodical loading for the slow-wave application and analysis

The ability to support a wave with a phase velocity much less than that of light is quite important for the traveling-wave-tube circuits. In a traveling-wave tube, efficient interaction between the electric beam and electromagnetic field is obtained only if the phase velocity is equal to the beam velocity. Since the beam velocity is often not greater than 10 to 20 percent of the speed of light [25]. Considerable slowing down of the electromagnetic wave by using periodic loading is required. It had been verified in the past that periodic structures are suitable to be used in traveling-wave tubes. The

rectangular waveguide is periodically loaded by thin capacitive or inductive irises [20]. By this way, the phase delay and phase advance are produced respectively.

There are also many analysis methods on the periodic structures. The propagation characteristics of the circular waveguide periodically loaded with dielectric disks are analyzed using the extension of the coupled-integral-equation technique [28]. The propagation constants are determined by the classical eigenvalues of a non-Hermitian matrix instead of a determinant. Hence, the computation time is significantly reduced. The transmission line matrix (TLM) method is applied to longitudinally periodic structures [29]. Phase wall derived from the Floquet's theorem [25] and the parameter estimation method for reducing the simulation time is used.

b) Periodically loaded structures for resonators and filters

The traditional combline resonator periodically loaded with two dimension (2-D) periodic ring structures in the inner rode of the resonator is demonstrated in [24]. The resonator with periodic loading keeps all the merits of the conventional combline resonator, at the same time, it achieves much higher Q-factor and better spurious characteristics than its counterparts. For the same size and material, the Q-factor of the conventional resonator (9938) is much less than that of combline resonator with loading (45530).

The periodic structure of the metal strip is used to realize the compact SAW (Surface Acoustic Wave) filters and diplexer with the performance which can meet the requirement of DECT (Digital Enhanced Cordless Telecommunication) system [26].

In [30], the microstrip (MS) transmission line is periodically loaded with MS ring or stepped-impedance hairpin resonators to realize the slow-wave filters, which demonstrate lower insertion loss than the parallel coupled filters. The MS line low pass filters (LPFs) with uniform circular ground defects, non-uniform circular ground defects and non-uniform defect rings on the ground with uniform, binomial and Chebyshev distribution of the circular pattern are investigated extensively in [31].

c) Periodically loaded structures for antenna

The loading technique has been proved useful technique in designing both the resonant and non-resonant antenna arrays [27]. For the resonant arrays typically used for airborne fire control applications, the impedance bandwidth becomes very narrow. When the number of the elements is increased, the increased bandwidth is of obvious benefit. For those non-resonant arrays typically used for radar applications, the impedance bandwidth of the overall arrays can be very broad without loading. However, the loading can improve the pattern width, since the reflections among sections of the array are reduced. Du and Gong [33] has reported that a planar inverted-F antenna (PIFA) with the periodic PBG structures on the ground plane. The loaded PIFA has almost the same operating bandwidth, antenna directivity and pattern characteristics, while the size is smaller than that of the traditional PIFA.

d) Periodically loaded structures on CPW

The coplanar waveguide (CPW) is an attractive transmission line for periodic

loading. One dimension (1-D) periodically loaded structures are reported in [32]. The periodic loading improves the stopband performance and suppress ripple near the cutoff frequency. The CPW periodic structures with less number of periods by using resonant elements on the lateral ground are used to improve the attenuation in the stopband [21]. The operating frequency of the periodic structures becomes tunable by using high permittivity material or polarizing diodes in defects [22]. Three periodically loaded structures and their lumped element equivalent circuit models are introduced in [23]. The reported low pass filters (LPFs) [23] with these loading structures has much smaller size than the traditional step width LPFs. At the same time, better performance in the filter stopband is also achieved.

In the following section, the CBCPW with capacitive periodic loading will be deduced. CBCPW becomes more and more attractive in the application of microwave integrated circuit (MIC), MMIC, RF-MEMS and other circuits technologies, partially due to its several advantages over the conventional microstrip (MS) lines [22]-[23]. These advantages include easy integration with active components, flexibility in changing the propagation constant and characteristic impedance by selecting different width ratios of the slot to strip, and low dispersion for both the propagation constant and the characteristic impedance of the CBCPW mode. So it is desirable to combine the advantages of CBCPW and periodic loading together to improve the performance and reduce the size of the circuits. Similar analysis in [25], but for CBCPW, is performed as a contrast to the analysis in Chapter 4. Periodically loading analysis in this chapter can not applicable to the gradually changed discontinuity. In Chapter 4, a

full-wave parameter extracting method is used for gradually loading with frequency-dependency, and a novel method of cell cascading with phase compensation is proposed to reduce the computation.

2.3 Capacitive Loaded CBCPW Analysis

To introduce a number of basic concepts, methods of analysis, and typical properties of periodic structures, we shall consider a simple example of capacitively loaded transmission line. Similar deduction is also suitable to the inductively loaded transmission line. For a physically smooth transmission line, the phase velocity is given by equation (2.19)

$$V_p = 1/\sqrt{LC_a} = 1/\sqrt{\mu_0 \epsilon_e \epsilon_0} \quad (2.19)$$

where ϵ_e is the effective relative dielectric constant of coplanar waveguide backed by a ground plane, ϵ_0 and μ_0 are electric permittivity and magnetic permeability in free space respectively. A significant reduction in the phase velocity can be achieved in a smooth line only by increasing ϵ_e . The disadvantage of the method is that the cross-sectional dimensions of the line must also be reduced in order to avoid the propagation of higher-order modes. The phase velocity cannot be decreased by increasing the shunt capacitance C_a of per unit length, since any change in the line configuration to increase the C_a automatically decreases the series inductance L of per unit length according to $LC_a = \mu_0 \epsilon_e \epsilon_0$. However, by removing the restriction that the line should be physically smooth, an effective increase in shunt capacitance per unit length can be achieved without a corresponding decrease in the series inductance L .

This means that the lumped shunt capacitance may be added at periodic intervals without offsetting the value L . If the spacing between the added lumped capacitors is smaller as compared with the wavelength, it may be anticipated that the line will appear to be electrically smooth, with a velocity shown in equation (2.20)

$$V_p = 1/\sqrt{L(C_a + C_0/d)} \quad (2.20)$$

where C_0/d is the amount of lumped capacitance added per unit length (A capacitor C_0 added at interval). The following analysis will verify this conclusion.

Figure 2.7 Thin diaphragm shunt capacitive loading of a CBCPW line

One method of obtaining shunt capacitive loading of a CBCPW transmission line is to introduce thin diaphragm at regular intervals as given in Figure 2.7. The fringing electric field in the vicinity of the diaphragm increases the local storage of the electric energy. Hence it may be considered, from the circuit viewpoint, as a new shunt capacitance. The local field can be described in terms of the incident, reflected, and transmitted TEM mode and a superposition of an infinite number of high-order modes. If the distance spacing $b-a$ is smaller as compared to the wavelength and the substrate thickness, the high-order modes are evanescent and will decay to negligible value at a distance of the order of $b-a$ away from the diaphragm in either direction. Only the

Center strip

d

dominant mode is considered and the leakage modes and the high-order modes are not considered.

Figure 2.8 The equivalent circuits for the unit cell and cascade cells

a) Equivalent circuit loaded unit cell b) cascade network of cells

The circuit analysis of a periodic structure involves constructing an equivalent network for a unit basic section or unit cell of the structure. This is followed by an analysis of determining the voltage and current waves that propagate along the network, which consists of an equivalent network of a basic section. The basic section is a shunt normalized susceptance $\bar{B}$ with a length $d/2$ of the transmission line on either side, as shown in Figure 2.8 a). Figure 2.8 b) illustrates the voltage-current relationships at the input and output of the n^{th} section in the infinitely long cascade connection.

The relationships between the input variables V_n, I_n and output variables V_{n+1}, I_{n+1} are readily found from ABCD matrix analysis. The V_n and I_n (amplitudes of the total voltage and current) are the sum of the contributions from the incident and reflected TEM waves at the terminal plane. The circuit for a unit cell may be broken down into three circuits in cascade, namely, a section of transmission line with the length of $d/2$

(electrical length $\theta/2 = K_0 d/2$), followed by a shunt susceptance $\bar{B}$, which in turn followed by another transmission line with length of $d/2$. The ABCD matrix for each of these individual networks is

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{\text{unit-section}} = \begin{bmatrix} \cos(\theta/2) & j \sin(\theta/2) \\ j \sin(\theta/2) & \cos(\theta/2) \end{bmatrix}_{\text{line}} \times \begin{bmatrix} 1 & 0 \\ j\bar{B} & 1 \end{bmatrix}_{\text{load}} \times \begin{bmatrix} \cos(\theta/2) & j \sin(\theta/2) \\ j \sin(\theta/2) & \cos(\theta/2) \end{bmatrix}_{\text{line}} \quad (2.21)$$

The transmission matrix for the unit cell is obtained by the chain rule, i.e., the product of the above three matrices. Hence, we have

$$\begin{bmatrix} V_n \\ I_n \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix} = \begin{bmatrix} \cos \theta - \frac{\bar{B}}{2} \sin \theta & j(\frac{\bar{B}}{2} \cos \theta + \sin \theta - \frac{\bar{B}}{2}) \\ j(\frac{\bar{B}}{2} \cos \theta + \sin \theta + \frac{\bar{B}}{2}) & \cos \theta - \frac{\bar{B}}{2} \sin \theta \end{bmatrix} \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix} \quad (2.22)$$

Note that $A=D$, which is always true for a symmetrical network, i.e. a symmetrical unit cell.

If the periodic structure is capable of supporting a propagation wave, it is necessary for the voltage and current at the $(n+1)^{\text{th}}$ terminal to be equal to the voltage and current at the n^{th} terminal, apart from the phase delay, due to a finite propagation time. Thus we assume that:

$$V_{n+1} = e^{-\gamma d} V_n \quad (2.23a)$$

$$I_{n+1} = e^{-\gamma d} I_n \quad (2.23b)$$

where $\gamma = \alpha + j\beta_n$ is the propagation constant for the periodic structure. In terms of the transmission matrix for a unit cell, we now have

$$\begin{bmatrix} V_n \\ I_n \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix} = e^{\gamma d} \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix} \quad (2.24a)$$

or

$$\left(\begin{bmatrix} A & B \\ C & D \end{bmatrix} - \begin{bmatrix} e^{\gamma d} & 0 \\ 0 & e^{\gamma d} \end{bmatrix} \right) \begin{bmatrix} V_{n+1} \\ I_{n+1} \end{bmatrix} = 0 \quad (2.24b)$$

This equation is a matrix eigenvalue equation for γ . A nontrivial solution for V_{n+1}, I_{n+1} exists only if the determinant vanishes, hence:

$$\begin{vmatrix} A - e^{\gamma d} & B \\ C & D - e^{\gamma d} \end{vmatrix} = AD + e^{2\gamma d} - (A + D)e^{\gamma d} - BC = 0 \quad (2.25)$$

For a reciprocal two port network the determinant $AD - BC$ of the matrix equals unity, so we can obtain

$$\cosh \gamma d = \frac{A + D}{2} \quad (2.26)$$

For the capacitively loaded CBCPW line, equation (2.26) together with equation (2.22) yields:

$$\cosh \gamma d = \cos \theta - \frac{\bar{B}}{2} \sin \theta \quad (2.27)$$

When $\left| \cos \theta - \frac{\bar{B}}{2} \sin \theta \right| < 1$, we must have $\gamma = j\beta$ and $\alpha = 0$, that is,

$$\cos \beta d = \cos \theta - \frac{\bar{B}}{2} \sin \theta \quad (2.28a)$$

When the right-hand side of equation (2.27) is greater than unity, $\gamma = \alpha$ and $\beta = 0$ so

$$\cosh \alpha d = \cos \theta - \frac{\bar{B}}{2} \sin \theta \quad (2.28b)$$

Finally, when the right-hand side of equation (2.25) is less than -1, we must have $\gamma d = j\pi + \alpha$, so that

$$\cosh \gamma d = \cosh(j\pi + \alpha d) = -\cosh \alpha d = \cos \theta - \frac{\bar{B}}{2} \sin \theta < -1 \quad (2.28c)$$

It is apparent that there will be frequency bands in which un-attenuated propagation can take place separately from frequency bands in which the wave is attenuated. Note

that propagations in both directions are possible since $-\gamma$ is provide another solution.

A detailed study of the passband and stopband characteristic is made by studying the characteristic of K_0 - β diagram. Now, we shall confine our attention to the low-limiting value of β_n . When $d \ll \lambda$, $\theta = k_0 d$ is small and $\beta_n d$ will then also be small. Replacing $\cos \theta$ by $1 - \theta^2 / 2$ and $\sin \theta$ by θ in equation (2.28a) gives

$$\cos \beta_n d \approx 1 - \frac{\beta_n^2 d^2}{2} = 1 - \frac{k_0^2 d^2}{2} - \frac{\bar{B} k_0 d}{2}$$

Using the relations

$$k_0^2 = \omega^2 \mu_0 \epsilon_0 \epsilon_e = \omega_0^2 L C_a \text{ and } \bar{B} = B / Y_0 = \omega C_0 (L / C)^{1/2}$$

where $\omega C_0 = B$ and $Z_0 = 1 / Y_0$ is the characteristic impedance of the transmission line, we obtain

$$\beta_n^2 = \omega^2 L C_a + \omega^2 L C_0 / d$$

hence

$$\beta_n = \omega \sqrt{L(C_a + C_0 / d)} \quad (2.29)$$

Therefore, we find that, at the low frequency, where $d \ll \lambda_0$, the loaded line behaves as an electrically smooth line with shunt capacitance $C_a + C_0 / d$ per unit length. The increase of β_n results in the reduction of the phase velocity by a factor k_0 / β_n .

Another important parameter in connection with periodic structures is the normalized characteristic impedance $\bar{Z}$ (due to the periodic loading, the characteristic impedance Z is no longer equal to that of the unloaded one with characteristic impedance Z_0) presented to the voltage and current waves at the reference terminal plane, i.e., input terminal of the unit cell. An expression for $\bar{Z}$ may be obtained from equation (2.24b) which can be written as

$$(A - e^{\gamma d})V_{n+1} = -BI_{n+1}$$

$$-CV_{n+1} = (D - e^{\gamma d})I_{n+1}$$

Hence

$$\frac{Z}{Z_0} = \bar{Z} = \frac{V_{n+1}}{I_{n+1}} = \frac{-B}{A - e^{\gamma d}} = -\frac{D - e^{\gamma d}}{C} \quad (2.30)$$

Replacing $2e^{\gamma d}$ by $A + D \pm [(A + D)^2 - 4]^{1/2}$ from equations (2.23a) and (2.23b), we obtain

$$\bar{Z}^{\pm} = \frac{2B}{D - A \pm \sqrt{(A + D)^2 - 4}} \quad (2.31a)$$

where the “+” and “-” sign refer to propagation in the +z and -z directions respectively. We are using the convention that the positive directions of V_n and I_n are those indicated in Figure 2.8. For a symmetrical network, $A=D$ and $AD-BC=1$, we have $A^2-1=BC$. In this case, equation (2.31a) reduces to

$$\bar{Z}^{\pm} = \frac{2B}{\pm \sqrt{4A^2 - 4}} = \pm \sqrt{\frac{B}{C}} \quad (2.31b)$$

In general, for a lossless structure, we have $\bar{Z}^- = -(\bar{Z}^+)^*$.

If unit cell is represented by a T network with parameters $\bar{Z}_{11}$, $\bar{Z}_{12}$ and $\bar{Z}_{22}$, by using the relations between the ABCD parameters and impedance, we can also show that

$$\cosh \gamma d = \frac{\bar{Z}_{11} + \bar{Z}_{22}}{2\bar{Z}_{12}} \quad (2.32)$$

$$\bar{Z} = \frac{\bar{Z}_{11} - \bar{Z}_{22}}{2} \pm \bar{Z}_{12} \sinh \gamma d \quad (2.33)$$

The waves that may propagate along a periodic structure are often called Bloch waves (analogy with quantum-mechanical electron waves propagating through a periodic

crystal lattice in a solid). The voltage and current at the n^{th} terminal plane will be denoted by $V_n^{\pm}$ and $I_n^{\pm}$ for the Bloch waves from now on instead of by the quantities V_n and I_n , where the “+” and “-” signs refer to Bloch waves propagating in the $+z$ direction and $-z$ direction. We shall also adopt the convention that the positive direction of current flow for Bloch waves is always in the $+z$ direction; thus $I^+ = \bar{Y}^+ \cdot V^+$ and $I^- = \bar{Y}^- \cdot V^-$, however, for a symmetrical structure such that $A=D$, we shall have $\bar{Y}^- = -\bar{Y}^+ = -(\bar{Z}^+)^{-1}$

If equation (2.31a) is used, we find that, for the loaded CBCPW line

$$\bar{Z} = \sqrt{\frac{\bar{B}}{C}} = \sqrt{\frac{2 \sin \theta + \bar{B} \cos \theta - \bar{B}}{2 \sin \theta + \bar{B} \cos \theta + \bar{B}}} \quad (2.32)$$

In the low frequency limit, where we can replace $\sin \theta$ by

$$\theta = k_0 d = \omega d / \sqrt{LC}$$

and $\cos \theta$ by 1, we obtain $\bar{Z} = \sqrt{\frac{2\theta}{2\theta + 2\bar{B}}} = \sqrt{\frac{C_a}{C_a + C_0/d}}$

$$Z = \bar{Z} \cdot Z_0 = \sqrt{\frac{L}{C_a + C_0/d}} \quad (2.33)$$

Again we see that, in the low-frequency limit, the loaded line is electrically smooth and the characteristic impedance is modified in the anticipated manner by effective increase in the shunt capacitance per unit length.

2.4 Conclusion

In this chapter, the concepts of filter and periodic loading, which may be used in the following chapters, are introduced. This introduction is helpful for understanding the following chapters. For example, the novel SEMCP filter topology, contrast to

filter topologies in section 2.1.3 and section 2.1.4, is proposed in Chapter 3 to achieve the quasi-elliptical response. The periodic loading analysis in section 2.3, which is good for understanding the loading principle, is not applicable to gradually changed periodic structures. The cell parameter-extraction in Chapter 4 is used to achieve frequency dependent inverters, contrast to the static inverter in section 2.1.2. A novel design method of cell cascading with phase compensation is proposed.

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CHAPTER 3

COMPACT FILTER DESIGN ON PCB

With today's trend of technologies advancement, compact sized and high-performance microwave filters and diplexers are highly demanded in many communication systems. Although microstrip (MS) planar filters and diplexers on PCBs are relatively large, they are still drawing much attention due to the advantages of low cost and easy fabrication. There are two fundamental ways to reduce the MS filter size: one is to decrease the dimensions by using the substrate with high dielectric constant, the other way is to optimize the resonator dimension and coupling topology of the filters. This chapter focuses on the second method.

Since resonators are fundamental elements and consume most of the area in the filters and diplexers, many studies have been done to investigate the size reduction of MS resonators. The hairpin filters [1]-[4] make progress in the size reduction from parallel-coupling structures to U-shape resonators. Further progress in the size reduction is made by the miniaturized hairpin resonator filter [4], where two arms of the U-shape MS structure are further folded to form a pair of closely coupled lines to enhance the capacitive nature of the open-end arms. The dual-mode ring resonator is one of the alternative means for a miniaturizing band pass filters (BPFs). Many studies have been performed to explore the mechanism of coupling between the dual modes [5]-[6]. However, from the viewpoint of the effective electric length of the resonator, the miniaturization in [2]-[6] is directly or indirectly based on the

half-wavelength resonator. Generally speaking, the resonator size is twice the size of a quarter-wavelength resonator. For filter miniaturization, the quarter-wavelength resonators are better than the half-wavelength resonators. For the quarter-wavelength filter [10]-[11], [23], the frequency of the second passband is around triple of the fundamental frequency, thus the quarter-wavelength filter has wider stopband as compared to the half-wavelength filter. On the other hand, to reduce the filter size, the coupling space between the adjacent resonators should be as small as possible. However, for narrow bandwidth and large stopband rejection BPFs, the coupling coefficient between the inter-stage resonators should be small. This means large space between the inter-stage resonators for the quarter-wavelength filter (for example, the interdigital filter [1]) must be kept. If the coupling space for the filter can be reduced (especially for the narrow bandwidth filter), it will help to reduce the total size of the filters. Finally, the transmission zero points (ZPs) outside of the passband help to improve the skirt selectivity and result in the further miniaturization of a BPF through the reduction of the filter orders [7]-[9].

In this chapter, the common via open-ground (OG) structure is investigated and used to realize BPFs and diplexer with minimized size and improved performance. It will also investigate how to perturb the magnetic coupling with different kinds of perturbation to achieve different coupling coefficients and to generate more ZPs in stopband. The tap-line OG filter is characterized and designed by using the physical scalable lumped circuit model. A filter and a diplexer with multiple ZPs based on the OG structure and source-load electric coupling are investigated and designed by the

full-wave EM simulation as well as the lumped circuit model, and then verified by practical experiment. A spurious suppression method is proposed for the quarter-wavelength resonator and filter. An OG filter with wide stopband and deep stopband rejection is implemented. A novel general SEMCP filter configuration and topology are proposed to realize the second and higher order BPFs with compact size, low insertion loss and additional multiple ZPs. The proposed ZP generation technique is an alternative method for cross-coupled filter to obtain the quasi-elliptic response using cascaded resonators.

This chapter is arranged as follows. The concept and advantages of the OG structure are introduced in Section 3.1. In Section 3.2, the OG structure is analyzed by using the even-odd mode analyses. The different coupling coefficients of the OG structure corresponding to four different kinds of perturbations are presented. In Section 3.3, an OG filter is designed and investigated by using a frequency-dependent scalable model. A U-shape OG filter with the source-load coupling is investigated by using the full-wave EM simulation as well as the lumped circuit model in Section 3.4. A closed-form expression is helpful to provide insight into the generation of ZPs in the frequency domain. An OG type diplexer is implemented by using U-shape OG filters. A new spurious suppression method for quarter-wavelength resonator and filter is proposed and investigated in Section 3.5. A filter with ultra-wide stopband up to $5f_0$ is implemented. In Section 3.6, the compact separate electric and magnetic coupling paths (SEMCP) filters are investigated by using odd- and even-mode analyses and designed by the full-wave EM simulation and finally implemented on PCB. The

general filter topology of the SEMCP filter is also introduced in this chapter.

3.1 Open-Ground Structure

Figure 3.1 Proposed OG configuration

The OG configuration as shown in Figure 3.1 comprises a section of open ended (OPD) half-wavelength transmission line with the inductive shunt element Z_t loaded at the ground reference (GR) plane in the middle of the OPD transmission line. The loaded inductive shunt element Z_t can be replaced by an inductor or MS short-ended stub with high characterize impedance and is explained later.

The resonance of an OPD MS half-wavelength resonator or its modified configuration occurs in odd-modes. This means that at resonance, the voltage at the centre of the MS line is its' minimum, while the voltages at both ends (OPD) of the lines have maximal values with opposite signs as illustrated in Figure 3.2 a). When the center of the OPD line is perfectly grounded, i.e. GR in Figure 3.2 b) is perfect grounded, the voltage at GR is equal to zero and the current flows in the ground for any of the operating resonant modes. Hence, some of the even-modes which require the GR to be non-zero voltage are suppressed completely. So the configuration can

operate with the fundamental mode and its odd modes (frequency is $(2n-1)f_1$ where n is positive integer) as shown in Figure 3.2 b) (where f_2 and f_3 are about 3 times and 5 times of f_1 , respectively). Under this condition, the half-wavelength MS resonator splits into two independent quarter-wavelength resonators for the fundamental mode which operates separately with no coupling.

a) without perturbation

b) with perturbation

Figure 3.2 The voltage distribution on the half-wavelength line

However, when the GR is not perfect (with imperfect ground (IG)), in most practical cases, the voltage in the GR will then not be zero. The two quarter-wavelength resonators are dependent on the voltage potential in GR and can be treated as a dual mode resonator compared with the half-wavelength OPD MS

resonator. Under this condition, the IG can be seen as a perturbation, which can be used to realize the coupling between the two resonators. The IG perturbations can be realized by using a lumped inductor or inductive short-ended MS stub. The advantages of using these OG structures are given as follows:

- 1) The circuit will be more compact: the resonators have the size of less than a quarter-wavelength. The shunt Z_t between the two resonators also contributes to the size reduction.
- 2) The reactance or susceptance slope parameters of quarter-wavelength resonator are half as large as that of the half-wavelength resonators [12].
- 3) Due to the shorter resonators and weaker couplings, the characteristics of the resonators are nearly lumped. The resonator can be modeled using lumped equivalent circuit. Miniaturization of the structure can be realized by folding it to a compact shape (See U-shape OG and SEMCP filters in the following).
- 4) The similar frequency response characteristics in both the fundamental frequency range and the harmonic frequency range can be used to generate more ZPs in the high rejection band (HRB). These ZPs can improve the skirt selectivity as well as the stopband width and rejection level.

The above advantages will be demonstrated in the following sections.

3.2 Coupling Characteristics of the OG Structure

Figure 3.3 Equivalent circuits of odd- and even- Modes

a) Even-mode circuit of OG structure b) Odd-mode circuit of OG structure

For the physical symmetrical OG structure as shown in Figure 3.1, the I/O ports are symmetrical along the MS line with the physical length of l from OPD to GR plane. The shunt IG is represented by an inductive lumped element Z_t (Z_t is positive and pure imaginary). The odd-even mode analyses [10] can be adopted to analyze this structure. The even- and odd- mode equivalent configurations are illustrated in Figures 3.3 a) and b), respectively. The OG structure resonates when its input admittance is zero for both even and odd modes [10].

$$Y_{even} = Y_{odd} = 0 \quad (3.1)$$

For the even mode case, the equivalent circuit representation of a half OG structure is illustrated in Figure 3.3 a) when the magnetic wall is applied in the GR plane. The normalized even mode input admittance is derived as

$$\overline{Y}_{even} = \frac{j \tan \theta_x^e (2\overline{Z}_t + j \tan \theta_{l-x}^e) + 1 + j 2\overline{Z}_t \tan \theta_{l-x}^e}{2\overline{Z}_t + j \tan \theta_{l-x}^e} \quad (3.2)$$

where

$$\overline{Y}_{even} = Y_{even} / Y_s, \overline{Z}_t = Z_t / Z_s \text{ and } Y_s = 1 / Z_s \text{ (} Z_s \text{ is the characteristic impedance of the}$$

line). $\theta_x^e = 2\pi f_{even}x / v_{pe}$ and $\theta_{l-x}^e = 2\pi f_{even}(l-x) / v_{pe}$ denote the even-mode electrical lengths of transmission lines with physical length x and $l-x$, respectively. v_{pe} is the even-mode phase velocity of the MS line. Hence, the even-mode resonant frequency is given by

$$(\theta_x^e + \theta_{l-x}^e) = \tan^{-1}\left(-\frac{1}{j2Z_t}\right) \quad (3.3)$$

$$f_{even} = v_{pe} \tan^{-1}(jZ_s / 2Z_t) / 2\pi l \quad (3.4)$$

For the odd mode case, the equivalent circuit representation of an OG resonator is illustrated in Figure 3.3 b), when the electric wall is applied in the GR plane. The normalized odd mode input admittance is derived as

$$\bar{Y}_{odd} = \frac{1 - \tan \theta_x^o \tan \theta_{l-x}^o}{j \tan \theta_{l-x}^o} \quad (3.5)$$

where $\bar{Y}_{odd} = Y_{odd} / Y_s$

$\theta_x^o = 2\pi f_{odd}x / v_{po}$ and $\theta_{l-x}^o = 2\pi f_{odd}(l-x) / v_{po}$ denote the odd-mode electrical length of the open-ended transmission lines with physical length x and of a section of transmission line with physical length $l-x$, respectively. v_{po} is the odd-mode phase velocity of MS line. Hence, the odd-mode resonant frequency is given by

$$\theta_x^o + \theta_{l-x}^o = \frac{\pi}{2} \quad (3.6)$$

$$f_{odd} = \frac{v_{po}}{4l} \quad (3.7)$$

For the uniform MS line, $v_{po} = v_{pe} = v_p$, where v_p is the phase velocity of a MS line. From the above odd- and even-mode analyses, it can be observed that the even-mode is controlled by the imperfect ground Z_t , while the odd-mode frequency is independent of the perturbation and remains unchanged. The perturbation element is

shared by two resonators and the coupling is different from that of mutual magnetic coupling through the space [11].

The center frequency of the BPF can be approximated by averaging the even- and odd- mode frequencies as

$$f_c = \frac{1}{2}(f_{even} + f_{odd}) = \frac{v_p}{4l} \left[\frac{1}{2} + \frac{1}{\pi} \tan^{-1}(jZ_s / 2Z_t) \right] \quad (3.8)$$

The coupling between two resonators is characterized by the coupling coefficient K [12] which can be computed from even- and odd- mode frequencies as

$$K = \frac{f_{odd}^2 - f_{even}^2}{f_{odd}^2 + f_{even}^2} = \frac{\pi^2 - 4(\tan^{-1}(jZ_s / 2Z_t))^2}{\pi^2 + 4(\tan^{-1}(jZ_s / 2Z_t))^2} \quad (3.9)$$

It is to be noted that when the length l is fixed, f_{odd} is also fixed. The coupling coefficient of K depends on the value of Z_t , which affects the resonant frequency of the even-mode. The coupling coefficient K increases and the filter frequency f_c decreases with respect to the perturbation element value Z_t as illustrated in Figure 3.4 for the case of $f_{odd} = 3\text{GHz}$. The numerical value of K can be made small, which is needed in a narrow-band filter design, or large as needed in a wide-band filter design. For the special case, when $Z_t = j0$ (i.e. perfectly ground), $\tan^{-1}(jZ_s / 2Z_t) = \frac{\pi}{2}$. Therefore, $K = 0$, which means no coupling exists. Two resonators with quarter-wavelength electrical length are operating separately with the same frequency. When $Z_t = j\infty$, $\tan^{-1}(jZ_s / 2Z_t) = 0$ and $K = 1$, which means no perturbation exists. Two resonators become one half-wavelength resonator.

Figure 3.4 Coupling coefficient and resonance frequency of OG structure

As given in Figure 3.1, both ends of the OG structure are OPDs. It is convenient to use the inductive type perturbation at the GR to realize Z_t . Four types of distributed perturbations are introduced in Figure 3.5. A short section of high characteristic impedance line grounded by via, as used in [12], is employed in Figure 3.5 b). In Figure 3.5 a), the via is directly on the trace at GR. In Figure 3.5 c), the balanced high impedance lines grounded by vias can minimize the transition interaction between the stubs and the transmission line in series. In Figure 3.5 d) structure, in order to get the coefficient of weak coupling, vias are connected in parallel to the GR plane. The approximate value of Z_t can be computed by using the equations given in Table 3.1. It should be noted that the multiple via holes in parallel to the GR plane become quite complex because of increasing number of the via holes. The parasitic capacitance and mutual magnetic coupling will be stronger with the increased number of via. Therefore the d) case in Table 3.1 can be used to estimate the approximate value of the total inductance.

Figure 3.5 OG structures with four kinds of perturbations

a) ILV b) OLV c) BOLV d) PV

TABLE 3.1 SUMMARY OF FOUR TYPES OF IGS

Type	Name	Z_t	approximation Z_t	K value
a)	In-line via (ILV)	Z_{via}	$j\omega L_{via}$	Small <0.05
b)	Off-line via (OLV)	$Z_c \frac{Z_{via} + jZ_c \tan \theta_c}{Zc + j2Z_{via} \tan \theta_c}$	$\frac{Z_c \omega l}{v_p} + j\omega L_{via}$	Very large >0.1
c)	Balance OLV (BOLV)	$Z_c \frac{Z_{via} + jZ_c \tan \theta_c}{2(Zc + j2Z_{via} \tan \theta_c)}$	$\frac{1}{2} \left(\frac{Z_c \omega l}{v_p} + j\omega L_{via} \right)$	Large >0.1
d)	Parallel Via (PV)	Z_{via} / n n number of via	$j\omega L_{via} / n$	Very small <0.05

For these four kinds of perturbations, via is a fundamental element. At DC or low operating frequency, via is considered to a perfect short. When the frequency is up to several GHz, the parasitic effect must be considered by the model as shown in Figure 3.6. When $H < 0.03\lambda_g$ (λ_g is the wavelength on the microstrip substrate), an accurate and relatively simple closed-form equation for the equivalent inductance is given in [13]:

Figure 3.6 Equivalent circuit of perfected conducting via

$$L_{via} = \frac{\mu_0}{2\pi} \left[H \cdot \ln \left(\frac{H + \sqrt{r^2 + H^2}}{r} \right) + \frac{3}{2} \left(r - \sqrt{r^2 + H^2} \right) \right] \quad (3.10)$$

where H is the thickness of the substrate and r is the radius of the via. The parasitic capacitance is the parallel plate capacitance between the via pad and the ground plane and can be approximated by equation [14]:

$$C = \frac{A\epsilon_r\epsilon_0}{H} \quad (F) \quad (3.11)$$

where A is the area of the top pad, ϵ_0 and ϵ_r are free space permittivity (8.85×10^{-12} Farads/meter) and relative permittivity, respectively. This capacitance is relatively small (<0.05 pF for our case). Coupling introduced by this capacitance will not dominate in the overall coupling of a single via [12].

Figure 3.7 Comparison between the coupling coefficients of OLV and BOLV

Figure 3.8 Comparison of the coupling coefficients of ILV and PV

By using equations (3.8), (3.9), (3.10) and approximation equations in table III.I, the coupling coefficients of four types of perturbation are illustrated in Figures 3.7 and 3.8. Figure 3.7 demonstrates the relationship between K and the length L_c of the shunt high impedance MS line. K mainly depends on the impedance and length of MS lines, rather than on the characteristics of the via. Figure 3.8 shows the relations between the radius of via and the coupling coefficients for PV type and ILV type perturbation. Actually, Figure 3.7 and Figure 3.8 demonstrate both cases with and without the parasite capacitor. However, the coupling coefficient curves are stacked for these two cases and no difference can be seen in the graph scale. Thus the simplified Z_t in Table 3.1 can be used.

3.3 Model-Based Design of Tap-Feed OG Filter

Because of the short resonators and weak couplings, the quarter-wavelength resonator is nearly lumped [11] and can be easily modeled as introduced before.

Full-wave EM based simulation is adopted in accurate design of the quarter-wavelength filter. However, full-wave method is at the expense of large computational effort. The synthesis or model methods are widely accepted in quarter-wavelength filter design and analysis. The traditional lumped LC models are only valid for frequency-invariant coupling and resonator [24]. In [4], the frequency-variant coupling is investigated by using a frequency-variant inverter. However, it is well known that characteristics of transmission line filter depend not only on the couplings but also on the resonating element itself, both of which are frequency-variant. So the frequency-variant characteristics of resonating elements and coupling still need to be considered for the accuracy of filter design.

Figure 3.9 Configuration and the model of the proposed filter

a) Configuration b) Proposed model

In this section, a tap-line feed second-order OG filter as shown in Figure 3.9 a), is characterized by frequency dependent lumped circuit model as shown in Figure 3.9 b). It is derived from transmission line network. The effects of MS open end are also considered in the model. By using the configuration, a filter can be directly designed using our proposed frequency-variant LC equivalent circuit model. The

frequency-variant characteristics of resonating elements and coupling in the filter are included in the frequency-variant LC circuit network. The LC model is helpful to understand the operating mechanism of the filter. The satisfactory agreement between the measurement and simulation demonstrates that the frequency-variant LC model-based design is valid for the design and analysis.

3.3.1 Scalable model

According to the transmission line theory, the input admittance of lossless open-end stub, with a characteristic admittance of $Y_s = 1/Z_s$, the propagation constant of β_s and physical length of l_1 as shown in Figure 3.9, is given by

$$Y_{in} = jY_s \tan(\beta_s(l_1 + \Delta l)) = j\omega C_o \quad (3.12)$$

where Δl is the equivalent length of the open end. At the open end of a MS line with width of W and substrate thickness of H , the field does not stop abruptly but extend slightly further due to the fringing field. This effect can be modeled as an equivalent transmission line with length of Δl [25]

$$\Delta l = hP_1P_3P_5 / P_4 \quad (3.13)$$

where

$$P_1 = 0.434907 \frac{\epsilon_{re}^{0.81} + 0.26(W/H)^{0.8544} + 0.236}{\epsilon_{re}^{0.81} - 0.189(W/H)^{0.8544} + 0.87}$$

$$P_2 = 1 + \frac{(W/H)^{0.371}}{2.35\epsilon_r + 1}$$

$$P_3 = 1 + \frac{0.5274 \tan^{-1} [0.084(W/H)^{1.9413 \cdot P_2}]}{\epsilon_{re}^{0.9236}}$$

$$P_4 = 1 + 0.037 \tan^{-1} [0.067(W/H)^{1.456}] \cdot \{6 - 5 \exp[0.036(1 - \epsilon_r)]\}$$

$$P_5 = 1 - 0.218 \exp(-7.5W/H)$$

For the range of $0.01 \leq W/H \leq 100$ and $\varepsilon_r \leq 128$, the accuracy is better than 0.2%.

The input admittance can be modeled by the admittance of the equivalent capacitance in (3.12), which can be calculated by

$$C_o = \frac{Y_s \tan(\beta_s(l_1 + \Delta l))}{\omega} \quad (F) \quad (3.14)$$

For a section of lossless transmission line with the length of l_2 , the transfer matrix is given by

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{s2} = \begin{bmatrix} \cos(\beta_s l_2) & Z_s \sin(\beta_s l_2) \\ Y_s \sin(\beta_s l_2) & \cos(\beta_s l_2) \end{bmatrix} \quad (3.15)$$

The above section MS line can also be modeled as an equivalent π -network (one inductor in series with two shunt capacitors on each side of the inductor). The equivalent inductance of L and capacitance of C_{s2} can be determined by

$$L = \frac{Z_s \sin(\beta_s l_2)}{\omega} \quad (H) \quad (3.16)$$

$$C_{s2} = \frac{1 - \cos(\beta_s l_2)}{\omega Z_s \sin(\beta_s l_2)} \quad (F) \quad (3.17)$$

The single ground via can be modeled as a lumped elements as introduced before.

The accurate and relative simple closed-form equation for the equivalent inductance is given by [13]

$$L_{via} = \frac{\mu_0}{2\pi} \left[H \cdot \ln \left(\frac{H + \sqrt{r^2 + H^2}}{r} \right) + \frac{3}{2} \left(r - \sqrt{r^2 + H^2} \right) \right] \quad (H) \quad (3.18)$$

where r is the radius of via.

From the above analyses, the structure in Figure 3.9 a) can be modeled as the lumped equivalent circuit in Figure 3.9 b). The capacitance of C_{op} is calculated by

$$C_{op} = C_o + C_{s2} \quad (3.19)$$

The transmission line characteristics as well as the open end effects are considered.

Thus under lossless condition, the microstrip line filter shown in Figure 3.9 a) can be physically modeled by the frequency-variant LC network given in Figure 3.9 b).

3.3.2 Filter design

The odd- and even-mode resonant frequencies of the resonator in the model of Figure 3.9 b) can be determined by using even- and odd-mode analyses [9]

$$\omega_{odd} = \frac{1}{\sqrt{LC_{op}}} \quad (3.20)$$

$$\omega_{even} = \frac{1}{\sqrt{(L + 2\Delta)C_{op}}} \quad (3.21)$$

where Δ can be calculated by

$$\Delta = \frac{L_{via}}{1 - \omega^2(2C_{s2}L_{via})} \quad (3.22)$$

The center frequency of the BPF can be approximated by averaging the even- and odd- mode frequencies as

$$\omega_c = \frac{1}{2}(\omega_{even} + \omega_{odd}) = \frac{1}{2} \left[\frac{1}{\sqrt{LC_{op}}} + \frac{1}{\sqrt{(L + 2\Delta)C_{op}}} \right] \quad (3.23)$$

The coupling between the two resonators is characterized by the coupling coefficient K [10], which can be computed by using the known odd- and even- mode frequencies

$$K = \frac{\omega_{odd}^2 - \omega_{even}^2}{\omega_{odd}^2 + \omega_{even}^2} = \frac{\Delta}{2L + \Delta} \quad (3.24)$$

For narrow band operation, the external Q -factor [13] of the tapped line

quarter-wavelength resonator can be approximately extracted from a circuit model

$$Q_e = \frac{\omega_c C_{op}}{G_s} \quad (3.25)$$

The required inter-stage coupling parameters for a second-order filter can be calculated by using the function in [11]

$$K = \frac{\pi BW}{4\sqrt{g_1 g_2}} \quad (3.26)$$

where g_1 and g_2 are the element values of the low pass filter prototype as introduced in Chapter 2. BW denotes the fractional bandwidth. The calculated inter-stage coupling coefficient is used to determine the shunt element impedance L_{via} as given in equations (3.18) and (3.24), by which suitable dimensions of the shunt element can be determined. The external Q-factor of resonator is determined by [11]

$$Q_e = \frac{g_0 g_1}{\Delta} = \frac{g_2 g_3}{\Delta} \quad (3.27)$$

The Q_e value is used to determine the position of a tap using equation (3.25).

The above model-based derivation is convenient for the initial design of the filter. To get the frequency response over the frequency range of interest, the two-port network parameters are used and the transfer matrix of the model in Figure 3.9 b) can be calculated by

$$\begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} = \frac{1}{s\Delta} \begin{bmatrix} 1 & 0 \\ sC_{op} & 1 \end{bmatrix} \begin{bmatrix} s(\Delta+L) & s^2 L(2\Delta+L) \\ 1 & s(\Delta+L) \end{bmatrix} \begin{bmatrix} 1 & 0 \\ sC_{op} & 1 \end{bmatrix} \quad (3.28)$$

and

$$T_{11} = T_{22} = \left((L + \Delta) + s^2 LC_{op}(L + 2\Delta) \right) / \Delta \quad (3.29)$$

$$T_{12} = (sL(L + 2\Delta)) / \Delta \quad (3.30)$$

$$T_{21} = \left(2sC_{op}(L + \Delta) + s^3LC_{op}^2(L + 2\Delta) + 1/s \right) / \Delta \quad (3.31)$$

The transmission and reflection characteristics can be determined by

$$S_{21} = \frac{2}{2T_{11} + T_{12}/Z_0 + T_{21}Z_0} \quad (3.32)$$

$$S_{11} = \frac{T_{12}/Z_0 - T_{21}Z_0}{2T_{11} + T_{12}/Z_0 + T_{21}Z_0} \quad (3.33)$$

where Z_0 is the termination impedance of two ports.

The model-based simulations are performed using Matlab 6.5. The results are shown in Figure 3.10, Curves a) and b) demonstrate the open end effects on the filter characteristics. The effects of the open ends can shift the operating frequency of the filter to a lower side. At the same time, Q_e increases with increasing of the equivalent capacitance C_{op} . From the illustration in Figure 3.10 and Figure 3.11, good agreement between the model-based results and the simulation results from commercial circuit simulator (AWR Microwave office 6.0) over the frequency range of interest is demonstrated. From our investigation, the frequency responses of the model-based results and EM results in second passband also show good agreement. The frequency-variant lumped LC model, the characteristics of which are different from the traditional frequency-invariant lumped LC model, can accurately describe the characteristics of the filter.

Figure 3.10 Comparison of model-based results with and without open end effects

$l_2=1.5$ mm, $l_1=12.15$ mm, $W_0=W_s=0.6$ mm, $r=0.3$ mm (refer to Figure 3.9a)

a) without open end effects

b) with open end effects

Figure 3.11 Comparison of the results with and without open-end effects

$l_2=1.5$ mm, $l_1=12.15$ mm, $W_0=W_s=0.6$ mm, $r=0.3$ mm

a) without open end effects

b) with open end effects

Note: Ground via is replaced by an inductance of 0.19nH using equation (3.13)

3.3.3 Results and discussion

The filter is realized on RT/duroid 6010 dielectric substrate with relative

dielectric constant of $\epsilon_r = 10.2$ and thickness of 0.635 mm. The filter is measured with HP8722ES vector network analyzer using 3.5 mm connector calibration kits. The universal substrate test fixtures WK-3001-B from Inter-continental Microwave Inc. are used for I/O port connection and calibration.

Figure 3.12 Comparison of the model-based results and experimental results

$$l_2 = 1.55 \text{ mm}, l_1 = 12.25 \text{ mm}, W_0 = W_s = 0.6 \text{ mm (refer to Figure 3.9a)}$$

In Figure 3.12, the model-based results are compared with the measured results. Both results are in good agreement in the passband as well as in the stopband. The measured specifications of the filter are: operating frequency 1.94 GHz, 3dB bandwidth 160 MHz, the insertion loss 1.75 dB. The stopband attenuation is larger than 30 dB from 2.12 GHz to 2.56 GHz. The rejection at the first zero point of 2.15 GHz and the second zero point of 2.4 GHz are 33 dB and 47 dB respectively. The photograph of the fabricated filter is given in Figure 3.13.

Figure 3.13 Photograph of the fabricated BPF

Figure 3.14 Comparison of the theory results with and without electric coupling

However, the measured results have two transmission ZPs in the high stopband, while the model based results can only see one deep ZP in the high stopband. The bandwidth of the model-based results is a little larger than that of the measured results in the high passband. The discrepancies are further investigated. It is found that the weak mutual coupling in space, which is mainly dominant by electric coupling, has not been considered in the model-based design or transmission line simulator. So the operating bandwidth found in the measurement varies slightly as compared to that of the simulation. The dominant magnetic coupling and the weak electric coupling in space cancel out in the upper stopband. Thus an additional zero point in the high stopband is generated. To demonstrate these effects, a small capacitor (cap) with

capacitance of 0.013pF is added between two open ends and the results are shown in Figure 3.14. The modeled cap, which represents weak electric coupling of two resonators through space, can shift down the odd-mode frequency. It also reduce the bandwidth and generate additional zero Z_{p1} .

3.4 U-shape Filter and Diplexer

Figure 3.15 Responses of the different type second-order BPFs

In a filter design, it will be very helpful if extra transmission Zps can be created without sacrificing the passband response [14]. For example, extra transmission zeros can be tuned to reject possible interferences and to improve the stopband rejection of the filter. Furthermore, a lower order filter, which has a smaller circuit size and lower insertion loss than a higher order filter, may meet the stopband rejection specifications with the help of extra transmission zeros. In this section, a novel filter configuration with compact size and extra controllable ZPs is designed. The rolloff or skirt selectivity [10] of the second-order BPF with ideal chebyshev response of curve a) of

Figure 3.15 can be improved by using the same filter order with source/load (S/L) coupling in curve b) of Figure 3.15. This improvement is however achieved at the expense of the stopband rejection. The frequency response as shown in curve c) of Figure 3.15 is that of our proposed U-shape OG BPF, which has the same order as curves a) and b), the same good skirt selectivity but more controllable finite transmission zero point (Z_{p3}) as compared to the two typical second-order filters of curves a) and b) in Figure 3.15. The U-shape OG BPF will be investigated and used to implement the diplexer in the following sections.

3.4.1 Filter characterizing

The proposed U-shape OG filter and filter topology are shown in Figure 3.16 a) and Figure 3.16 b), respectively. In Figure 3.16 a), the inter-stage coupling is dominated by the magnetic couplings of the inductive perturbation in the GR as analyzed before. The signal from the input port P_1 is mainly coupling through the coupling gap S_1 to the resonator. At the same time, the direct signal path from P_1 to P_2

through the electric coupling gap S_2 exists. The weak electric coupling and the dominant magnetic coupling between the inter-stage resonators makes it generate the additional finite transmission ZPs in the frequency response of the filter. The required parameters for the filter function are given in [11]. For a second-order filter case, the equation (3.26) can be used to determine the coupling coefficient. $g_0=1.0$, $g_1=1.4142$, $g_2=1.4142$ and $g_3=1.0$ are selected in the following designs. The coupling coefficient is used to determine the shunt element impedance Z_t as given in equations (3.8) and (3.9). A suitable shunt type in Figure 3.5 can be chosen.

Figure 3.17 The external Q -factor of a resonator with respect to length L_1

The Q_e value is dependent on the resonator, coupling gap S_1 and the coupling line length L_1 . The EM software IE3D is then used to design and optimize the dimensions. Using EM simulation given in [10], the relationship between the Q_e and coupling length L_1 can be determined as shown in Figure 3.17.

Figure 3.18 OG filters with 6% bandwidth

$$S_1=0.1\text{mm}, L_1=2.95\text{mm}, L_2=5.69\text{mm}, W=0.61\text{mm}, R_{\text{via}}=0.4\text{mm}$$

The filters using ILV, with fractional bandwidth of 6% ($K=0.03$, $Q_e=21.8$, $f_c=3.53$ GHz), and the filters using OLV and BOLV, with fractional bandwidth of 18% ($K=0.1$, $Q_e=7.86$, $f_c=3.32$ GHz), are designed. The initial dimensions are chosen using the above mentioned design procedure. The optimized results are illustrated in Figure 3.18 and Figure 3.19, respectively. For the same length of L_1+L_2 , the bandwidth and center frequency of the two types of filters can be changed by using different dimensions or perturbation types of IG. As shown in Figure 3.18 and Figure 3.19, when the source to load coupling gap of S_2 is larger than 3mm, the zero point at high rejection band (HRB) is shifted up to harmonic frequency, while the ZP at the low rejection band (LRB) is shifted down towards DC. With the reduction gap width of S_2 , the ZPs at HRB and LRB are shifted closer to the center frequency of the filter. The skirt selectivity is improved but the stopband rejection is degenerated. It can also be

seen that the U-shape OG filter can generate an additional ZP at HRB.

Figure 3.19 OG filter with 18% bandwidth

Dimensions: $S_1=0.1$ mm, $L_1=7.2$ mm, $L_2=1.44$ mm, $W=0.61$ mm, $W_c=0.1$ mm, $R_{via}=0.2$ mm: BOLV1: $L_c=0.39$ mm, $S_2=3$ mm; BOLV2: $L_c=0.39$ mm, $S_2=0.81$ mm; OLV: $L_c=0.1$ mm, $S_2=3$ mm; BOLV3: $S_2=1.2$ mm, $L_c=0.39$ mm

3.4.2 Analytical modeling

Figure 3.20 The equivalent circuit model

When $j0 < Zt < j\infty$, OG structures behave as two coupled quarter-wavelength resonators with inductive loading. Each resonator has a parallel equivalent resonator model. Figure 3.20 illustrates the equivalent circuit of the OG filters, which is helpful

to understand the mechanism of transmission zero and pole positions in the frequency response. Δ is equivalent inductance calculated from Equation (3.22) of section 3.3. The direct electric coupling from the source to the load is modeled by the mutual coupling capacitance C_m .

From odd- and even- mode analyses, the odd and even-mode angular frequencies of the two resonators with shunt inductor Δ are $1/\sqrt{LC_{og}}$ and $1/\sqrt{(L+2\Delta)C_{og}}$ respectively. We use Matlab 6.5 to analyze the transmission and rejection characteristics of the lumped model. The results are shown in Figure 3.21. The derived HRB ZP frequency and LRB ZP frequency can be calculated by

$$S_{HRB,ZP} = \sqrt{\frac{(b+ag-e-hd) - \sqrt{(b+ag-e-hd)^2 + 8C_m(he-bg)}}{2(he-bg)}} \quad (3.34)$$

$$S_{LRB,ZP} = \sqrt{\frac{(b+ag-e-hd) + \sqrt{(b+ag-e-hd)^2 + 8C_m(he-bg)}}{2(he-bg)}} \quad (3.35)$$

If two real transmission ZPs exist, we have

$$(b+ag-e-hd)^2 + 8C_m(he-bg) \geq 0 \quad \text{and} \quad he-bg < 0$$

where

$$a = 2C_m + C_c$$

$$b = (2LC_mC_c + 2LC_mC_{og} + LC_cC_{og})$$

$$h = L(C_c + C_{og})$$

$$d = C_c$$

$$e = C_cC_{og}(L+2\Delta)$$

$$g = (C_{og} + C_c)(L+2\Delta)$$

Figure 3.21 Response of equivalent circuit of the OG Filter

Figure 3.22 The ZP positions in the spectrum versus coupling capacitance C_m

Figure 3.23 The relationship among the transmission Zps, C_m and Δ

In Figure 3.22 and Figure 3.23, it can be seen that for the case of the narrow band and dominant magnetic coupling, an increasing C_m leads to a closing of the transmission zeros to the center frequency of the BPF both in HRB and LRB. Thus the skirt selectivity of the filter has improved as shown in Figure 3.21. However, the increased C_m also reduces the rejection in the stopband. As illustrated in Figure 3.22 and Figure 3.23, when C_m (pF) $\ll \Delta$ (nH), a slight change of C_m leads to a large shift of the ZP position in the spectrum. When the C_m (pF) value is close to Δ (nH), the position of the transmission zeros almost remains constant. But the rejection in the stopband becomes worse. In general, ZPs are dependent on the ratio of C_m to Δ . That means ZPs depend on the ratio of the electric coupling strength to the magnetic coupling strength. The larger the ratio, the farther the distance from ZPs to the center frequency of the filter becomes. On the other hand, the decreasing value of C_m increases the stopband rejection, but at the same time, it decreases the roll-off besides the passband due to the shift of the ZPs. Therefore a tradeoff should be made between the skirt selectivity beside the passband and the rejection in the stopband. These equivalent circuit characteristics are helpful for the OG filter analysis. As shown in Figure 3.18 and Figure 3.19, the smaller the coupling gap S_2 , the closer the ZPs to the filter center frequency.

It is noted that there are differences between the response of the U-shape OG filter and its lumped equivalent circuit at the frequency range above HRB ZP. This is because half of the OG structure behaves as a simple lumped resonant circuit in the vicinity of a particular resonant frequency. It is in reality a much more complicated

network having infinite number of anti-resonance frequencies (first three as illustrated in Figure 3.1 b) in section 3.1) which occur when resonator is an odd multiple of a quarter-wavelength. Thus the exact equivalent circuit would consist of infinite number of resonant lumped circuits coupled together. However, our interested frequency is below the first harmonic resonance frequency. For the first harmonic (the case of f_2 as shown in Figure 3.1 b) in section 3.1), the circuit still has the similar models, but the element value for the resonator is about one third of that for the case of f_1 . For the harmonics, there are also two signal paths, an electric coupling path through C_m and a magnetic coupling path through Δ . Therefore, for the harmonic frequency of f_2 , the filter has similar transmission ZPs as for the fundamental frequency of f_1 . However, the coupling coefficient K becomes larger for the case of f_2 due to smaller equivalent element values L and C_{og} in the resonator model. Thus for the first harmonic frequency of f_2 , the ZP in LRB changes more quickly with the change of C_m than that for the fundamental frequency of f_1 . This characteristic brings great advantages to improve the skirt selectivity and the stopband characteristics simultaneously. Two ZPs in HRB change according to C_m with different directions (i.e. one shifts upwards to increase the skirt selectivity and the other shifts downwards to increase the stopband bandwidth). The EM results of the U-shape OG filters in Figure 3.18 and Figure 3.19 show that different S_2 (related to C_m in equivalent circuit) leads to different ZP positions and stopband rejection.

3.4.3 Filter and diplexer

The filter and diplexer are realized on RT/duroid 6010 dielectric substrate with $\epsilon_r = 10.2$ and thickness of 0.635 mm. A filter is designed at 3.53 GHz with 6% bandwidth and dimensions of $S_1=0.1$ mm, $L_1=2.95$ mm, $L_2=5.64$ mm, $W=0.605$ mm, $R_{via}=0.4$ mm and $S_2=0.8$ mm. The optimized EM results and the lumped model results are compared in Figure 3.24. The model-based results follows well with EM results below the first ZP with the frequency of 4.2 GHz at HRB. To accurately design and model in wide bandwidth, the scalable model, which includes dispersion effect, source coupling as well as harmonic, is needed. The theoretical results from the commercial EM software IE3D are compared to the experimental results and shown in Figure 3.25. The measured insertion loss of 1.65 dB is smaller than the predicted 2.06 dB. Also the measured bandwidth is slight larger than predicted.

Figure 3.24 The EM results and the lumped model results.

The element values of the lumped circuit model: $L=2.1$ nH, $C=0.73$ pF, $C_c=0.2$ pF, $L_{via}=0.158$ nH, $C_m=0.025$ pF

Figure 3.25 Comparison between EM simulated results from IE3D and the measurement

To investigate this discrepancy, another commercial full-wave 3D EM software HFSS is used to simulate the filter. The results from HFSS and measurement are compared in Figure 3.26. It has shown clearly that a better agreement to the measurement is obtained than with IE3D. By comparing the three results in Figure 3.25 and Figure 3.26, the simulation results of the filter with IE3D have narrower bandwidth and higher operating frequency than the results with HFSS or from the measurement. Since the bandwidth is mainly determined by inter-stage coupling inverter, and the operating frequency is also affected by the shunt inductor which performs the loading for the OG resonators, the larger the inductance of the shunt inductor, the larger the bandwidth and the smaller the operating frequency. Therefore, it can be concluded that the calculated equivalent inductance of the common ground via hole is smaller than that with HFSS or from the measurement. The larger practical equivalent inductance L_{via} increases the coupling coefficient, which can increase the

bandwidth and reduce the insertion loss. A small discrepancy between the measured results and the simulation results with HFSS above the second ZP is mainly contributed by the process of the fabrication. It can not be fabricated as accurate as that of the metal trace and gap. The second-order filter has good skirt selectivity characteristics as well as stopband rejection. The filter has become more compact than the comb filter with a similar frequency response. The realized filter is only 7.6 mm × 3.4 mm, as illustrated in Figure 3.27.

Figure 3.26 Comparison between the EM simulated results from HFSS and the measurement

Figure 3.27 Photograph of the filter

For diplexers and multiplexers, the good skirt selectivity and stopband rejection are rather important especially for rejection of the adjacent channel. The generation and proper placement of finite transmission ZPs is an effective way to achieve good selectivity and this measure is widely adopted in communication systems such as transponders and transceivers of wireless communication. Since the U-shape OG BPFs have multiple finite transmission ZPs, they can be used to build up diplexer with good stopband characteristics.

Figure 3.28 The configuration of the diplexer

To design the OG microstrip diplexer, two OG BPFs working at two different frequencies of 3.1GHz (BPF1) and 3.5 GHz (BPF2) with fractional bandwidth of 5.5% and 6%, respectively, are designed separately (see the configuration shown in Figure 3.28). The transmission ZPs of the filters are properly designed to obtain good roll-off besides the passband and to achieve good rejection in the higher stopband. Then the MS T-junction is used as the connector between input ports of BPF1 and BPF2. It is convenient to choose the MS line with a 50Ω characteristic impedance connected between filter input ports and the T-junction. However, the length of these lines must be designed carefully without deteriorating the performance of each filter. Lengths $TL1$ and $TL2$ of the feed lines are adjusted to represent an open circuit at the

T-junction at the center frequencies of BPF2 and BPF1 of the diplexer, respectively.

This is accomplished by setting the transmission line lengths [15] $TL1$ and $TL2$ according to the approximation given by equations (3.36) and (3.37)

$$TL_1 \approx n\lambda_2/2 \quad (3.36)$$

$$TL_2 \approx n\lambda_1/2 \quad (3.37)$$

where $\lambda_1 = c/(f_1\sqrt{\epsilon_{\text{reff}}})$ and $\lambda_2 = c/(f_2\sqrt{\epsilon_{\text{reff}}})$ are the effective wavelengths of the MS lines in front of BPF1 and BPF2 at frequencies f_1 and f_2 , respectively. ϵ_{reff} is the effective dielectric constant of the MS line and c is the speed of the light in free space.

In order to obtain a smooth response in the rejection band and maintain a compact size for the diplexer, it is desirable to choose the feed line length of $TL1$ and $TL2$ as short as possible (integer $n=1$ in equations (3.36) and (3.37) is chosen). The initial diplexer results are optimized by using IE3D software. BPF1 and BPF2 are designed using the filter structures in Figure 3.16 a). The dimensions of the BPF1 operating in 3.1 GHz are $L_1=3.05$ mm, $L_2=6.05$ mm, $S_1=0.1$ mm, $W=0.605$ mm, $S_2=0.8$ mm, $R_{\text{via}}=0.4$ mm. The dimensions of BPF2 operating in 3.5 GHz are $S_1=0.1$ mm, $L_1=2.95$ mm, $L_2=5.64$ mm, $W=0.605$ mm, $R_{\text{via}}=0.4$ mm, $S_2=0.8$ mm. $TL1=16.7$ mm and $TL2=16$ mm. The simulated and measured results are shown in Figure 3.29 and Figure 3.30, respectively. A similar discrepancy between the simulation and measurement in bandwidth exists as discussed before. The advantages of good skirt selectivity as well as good stopband rejection are demonstrated. For each channel, two finite ZPs are generated in the HRB and one ZP lies in LRB. The simulated channel isolation between channel 1 and 2 is higher than 21.7 dB. The measured insertion loss for each

channel is around 2.6 dB and the measured isolation between two channels is larger than 15.4 dB. The dimension of the diplexer is shown in Figure 3.31 and is only 20 mm X 30 mm (excluding the SMA connectors).

Figure 3.29 Simulation results of the diplexer

Figure 3.30 Measured results of the diplexer

Figure 3.31 Photographs of the diplexer

3.5 Spurious Suppression

The MS BPFs with wide spurious-free stopband are highly demanded in communication systems especially for high quality communication systems, where receivers with very good band selectivity are necessary. Therefore, many efforts have been made to suppress or eliminate the harmonic responses of the resonator to improve the out-of-band characteristics. The traditional BPF using a half-wavelength resonator suffers from a poor skirt selectivity and spurious response from the multiple fundamental frequency f_0 . The stepped-impedance resonator (SIR) [16]-[19], which is formed by two or more lines with different characteristic impedances, is widely adopted to suppress the spurious responses. However, the step ratio, which mainly determines the stopband width of a SIR filter, is limited by the minimum trace width of the fabrication process as well as large metal loss of the thin metal traces and radiation loss due to the abrupt discontinuity of the step. In this part, a novel extension technique of filter stopband width by properly designing the I/O feeder is proposed in following.

Figure 3.32 Quarter-wavelength resonators with different feeder types

a) end-coupling b) tap feeder c) parallel coupling d) “T” feeder

As introduced before, the quarter-wavelength resonator [10]-[11] is half the size compared to the half-wavelength resonator, and the first spurious frequency of the quarter-wavelength resonator appears at $3f_0$. Thus the filter with quarter-wavelength resonators has a wider stopband compared with the filter with half-wavelength resonators. The I/O feeders are always needed for the filters. The typical feeder (or coupling) types for quarter-wavelength resonators with uniform width of the metal trace are shown in Figure 3.32. The coupling in Figure 3.32 a) is called end coupling and is used for narrow band end-coupled BPFs. Feeder types in Figure 3.32 b) and Figure 3.32 c) are widely used in interdigital filters and combline filters [10]. The first spurious frequency of the resonators with uniform width of the metal trace under these feeding conditions appears in $3f_0$ [11]. A spurious suppression technique is proposed and used to extend the stopband width of the quarter-wavelength filter by using proper dimensions of the “T” type feeders [20] as shown in Figure 3.32 d). A

BPF example using OG structure and “T” type feeder has been designed in following. The narrow bandwidth BPF with low insertion loss, wide stopband of up to $5f_0$ and good stopband rejection up to 54 dB (see Figure 3.38) in the upper stopband, confirms the usefulness of the spurious suppression technique using “T” feeders.

3.5.1 Characteristics of resonator

Figure 3.33 Configuration for characterizing the resonator with “T” feeder

The quarter-wavelength resonator and the “T” feeder are shown in Figure 3.32 d). In order to investigate the coupling characteristic of the “T” type feeder, it is convenient to add another port, I/O port 2, as shown in Figure 3.33, to investigate the transmission characteristics of the resonator with two loads. To neglect the output effects on the characteristics of the resonator and the “T” feeder, the larger gap size of S_2 between the resonator and the open end of the feed line at I/O port 2 is set (for this case $S_2=1$ mm).

Both the diameter of the ground via and the length $(l_1 + l_2)$ of the resonator are fixed. It was found that the length of the “T” type feeder not only affect the transmission characteristics at the fundamental frequency of the resonator but also the

spurious characteristics at $3f_0$. The three different dimensions of the feeder structure in Figure 3.33 are investigated. The fundamental frequency response and the maximum harmonic responses at $3f_0$ are listed in Table 3.2. The transmission responses of the resonator under the three conditions are compared in Figure 3.34. It is clear that properly choosing of the lengths l_1 can suppress the spurious frequency at $3f_0$. The ratio of length l_1 to length l_2 for suppression of $3f_0$ is about 2:1. Under this condition, the harmonic $3f_0$ of the resonator is approximately equal to the fundamental resonant frequency of the open-ended MS line with width W_1 and length l_1 at the end of the “T” type feeder.

TABLE 3.2 COMPARING FEEDER EFFECTS ON RESONATOR

Type	l_1 (mm)	l_2 (mm)	W_1 (mm)	S_1 (mm)	IL in f_0 (dB)	Spurious frequency at $3f_0$ (dB)
A)	12	1.8	0.2	0.1	25.2	-18.2
B)	5	8.8	0.2	0.1	28.4	-18.6
C)	9.2	4.2	0.2	0.1	28	-33.7

Figure 3.34 Comparison of the transmission under different feeder dimensions

The mechanism of the spurious suppression of the “T” feeder is now investigated. At operating frequency f_0 of the quarter-wavelength resonator, the electrical length of l_1 is about one sixth wavelength. The open-ended section line (with width of W_l and length of l_1) of the “T” feeder can be treated as a coupling capacitor [22] connected at tap point tp in Figure 3.33. While at $3f_0$ of the quarter-wavelength resonator, the open-ended section line has the electrical length of about a half-wavelength. Since the tap point is at the middle of the open-ended section line, its input impedance at the tap point is zero. The reflection coefficient becomes minus one and the loaded quality factor is infinite. The section line behaves as a band-stop resonator, and the attenuation at $3f_0$ of the quarter-wavelength resonator is obtained. This characteristic can be used to extend the stopband width of the filters as demonstrated in the following filter example.

3.5.2 Filter design

The configuration of a second-order BPF is shown in Figure 3.35. The ground via at the symmetrical GR (ground reference) plane is used as the K inverter providing the inter-stage coupling for the BPF. The center frequency of the BPF can be approximately determined by using equations (3.8) and (3.9).

Figure 3.35 Configuration of the filter with spurious suppression

As shown in Figure 3.35, the stopband rejection is mainly determined by the dimension of the ground via which can lead to different coupling coefficients as given by equations (3.8) and (3.9). Since the substrate thickness is fixed, the larger the diameter of the via hole, the smaller the magnetic coupling coefficient and also the better the stopband rejection. The filter is initially simulated by circuit models in circuit environment. By comparing the filter performance with and without T-junction using the circuit model, it is found that the T-junction effects on the filter performance are quite small. The filters are designed on RT/duroid 6010 dielectric substrate. The initial dimensions can be optimized using IE3D software. The transmission characteristics of these three BPFs, corresponding to the dimensions given in Figure 3.35 and via diameter of 0.2 mm, are compared and shown in Figure 3.36. It can be seen that the suppression of the harmonic at $3f_0$ for curve C) of Figure 3.36 is 25 dB larger than that for curves A) and B).

Figure 3.36 Comparison of the theory and experiment results of the proposed filter

A) $l_1=12$ mm, $l_2=1.8$ mm, $W_l=0.2$ mm, $S_l=0.1$ mm B) $l_1=5$ mm, $l_2=8.8$ mm, $W_l=0.2$ mm, $S_l=0.1$ mm C) $l_1=9.2$ mm, $l_2=4.6$ mm, $W_l=0.2$ mm, $S_l=0.1$ mm

3.5.3 Experiments and results

The predicted and measured results of the filter under condition C) of Figure 3.36 are compared in Figure 3.37 and Figure 3.38. The simulated and measured return losses are better than -25 dB. The centre frequency is 1.87 GHz, the -3 dB bandwidth is 160 MHz and the insertion loss is 1.7 dB.

Figure 3.37 Passband performance of proposed filter C) of Figure 3.36

In Figure 3.38, the stopband characteristics of the predicted and the measured results are illustrated. From 3 GHz to 5.7 GHz, the stopband rejection is larger than 41 dB. From DC to 1.1 GHz, the rejection is larger than 51.5 dB. In the bandwidth of $3.5f_0$ (e.g. from 2.15 GHz to 8.68 GHz), the rejection is larger than 20 dB. The photograph of the proposed filter is shown in Figure 3.39. The board size is 10 mm by 40 mm. This filter demonstrates better stopband performance as compared to the traditional BPF with the same order. Firstly, the large stopband rejection has been achieved due to the common via ground, which is approximately a small lumped inductor with less distribution effect as compared to the coupled transmission lines. Secondly, the upper stopband of the proposed filter is extended up to $5.1f_0$. This spurious suppression method, which uses the different behavior of the open-ended section line at different frequencies (i.e. short open-ended line can be modeled as a capacitor at f_0 and as a bandstop resonator at $3f_0$) in I/O section topology, is different from the method in [19].

Figure 3.38 Stopband performance of proposed filter C) of Figure 3.36

Figure 3.39 The enlarged photograph of the filter with spurious suppression

3.6 A Compact Size Coupling Controllable SEMCP Filter

In order to reduce the size of the planar filter without sacrificing the performance, many techniques have been reported in the literature [1]-[13]. For instance, instead of using a dual mode half-wavelength resonator [1]-[4] or ring resonator [5]-[7], a quarter-wavelength resonator has been used [27]-[29], which can also provide a wider stopband. It is desirable to design a filter that additional transmission ZPs can be generated without degenerating the passband response [15], [24], [28]-[29]. The additional ZPs of the filter can be adjusted to reject possible interferences and to improve the stopband rejection. Thus a low-order filter with the help of the additional ZPs can meet the stopband requirements which are usually achieved by the high-order filters. It is also known that the low-order filters generally have smaller size and lower insertion losses compared with the high-order filters. To generate more expected ZPs, the coupling mechanism must be investigated clearly. Many papers in the literature already addressed the coupling mechanism. The source-load coupling or cross coupled topologies of elliptical filters are widely adopted [27], [29], [31].

The general coupling matrix, introduced in Chapter 2, is used to represent the coupling relationship for filter synthesizing [30]-[35]. It can be seen that only one

coupling parameter exists between any two adjacent resonators in the coupling matrix. For the inter-stage coupling of the quarter-wavelength filter, the inter-stage coupling in these reports can be classified as electric coupling, magnetic coupling or mixed coupling (both the electric and magnetic coupling coexist and can not be separated in space due to the distributed effect). The inter-stage coupling of the most reported quarter-wavelength filters has one physical coupling path between two adjacent resonators with one of these three coupling types.

In this part, new general filter topology and configuration, which provides two tunable separated electric and magnetic coupling paths (SEMCPs), are investigated. To the best of our knowledge, the work, which had been done to investigate two quarter-wavelength resonators with two simultaneously controllable and separated coupling paths, is not reported before. Based on the topology, the OG structure is modified to a compact configuration, and the inter-stage coupling can be well controlled through separately controlling the two coupling paths. At the same time, additional ZPs can be generated in either the lower or the upper stopband near the passband by choosing different dominant coupling types in two coupling paths. The rolloff (or skirt selectivity) of the filters can be improved and the sizes of the filters can also be reduced. This ZP generation mechanism of SEMCP is different from that of source-load coupling or cross coupled topology of elliptical filters [28]-[30]. The electric and magnetic couplings in SEMCP can not be represented separately by general coupling matrix and topology in [31]-[34], [36], where only one coupling parameter exists between any two adjacent resonators. Second-order SEMCP filters

are analyzed by odd- and even-mode analyses, designed by full-wave EM simulation and verified by the experiment. The second-order SEMCP filter configurations are also extended to high-order filter and examples of miniaturized fourth-order SEMCP filters are designed and implemented.

3.6.1 Analysis of OG filters with SEMCP

Figure 3.40 a) shows the configuration of a second-order SEMCP filter loaded with lumped coupling elements (L_m and C_m). Two sections of transmission lines are connected in series by a lumped capacitor C_m , which is called an electric coupling path. And a lumped inductor L_m is shunt at the other two ends of two transmission lines, which is called a magnetic coupling path. The filter topology and the coupling relationship are illustrated in the filter topology shown in Figure 3.40 b).

Figure 3.40 Proposed second-order filter configuration and the topology

a) filter configuration with lumped coupling elements b) filter topology

In Figure 3.40 a), the I/O ports and the physical structure are symmetrical with respect to the reference plane AA'. An odd- and even- mode analyses [10] can be adopted to analyze this structure. From the mechanism of the resonance, the resonator resonates, when its input admittance is zero for both even- and odd- modes, i.e.

$$Y_{even} = Y_{odd} = 0 \quad (3.38)$$

For the even-mode case, the equivalent circuit representation of the resonator when the magnetic wall is applied at the AA' plane is demonstrated in Figure 3.41.

The even-mode input admittance is derived as

Figure 3.41 The equivalent even-mode configuration

$$Y_{even} = jY_c \left[\tan \beta_e l_1 + \frac{2\omega_e L_m Y_c \tan \beta_e l_2 - 1}{2Y_c \omega_e L_m + \tan \beta_e l_2} \right] \quad (3.39)$$

where β_e is the propagation constant at the even-mode resonance angular frequency of ω_e and $Z_c = 1/Y_c$ is the characteristic impedance of the resonator.

Using equation (3.38) and equation (3.39), we have

$$\frac{2L_m \omega_e}{Z_c} = \frac{1}{\tan \beta_e (l_1 + l_2)} \quad (3.40)$$

For a small L_m , $l_1 + l_2 \approx \lambda_g / 4$, the right hand side of equation (3.40) can be expanded

by using Taylor expansion at $l_1 + l_2 = \lambda_g / 4$. By omitting the high-order terms, equation (3.40) can be rewritten as

$$\frac{2L_m\omega_e}{Z_c} + \beta_e(l_1 + l_2) = \frac{\pi}{2} \quad (3.41)$$

Since

$$\beta_e = \frac{\omega_e \sqrt{\epsilon_{re}}}{c}$$

where c is the speed of light in the free space,

the even-mode resonant angular frequency is given by

$$\omega_e = \frac{\pi}{2(2L_m Y_c + A)} \quad (3.42)$$

where

$$A = \frac{\sqrt{\epsilon_{re}}(l_1 + l_2)}{c}$$

Figure 3.42 The equivalent odd-mode configuration

For the odd-mode case, the equivalent circuit representation of the resonator when the electric wall is applied at the AA' plane is illustrated in Figure 3.42. The odd-mode input admittance is

$$Y_{odd} = jY_c \left[\frac{2\omega_o C_m + Y_c \tan \beta_o l_1}{Y_c - 2\omega_o C_m \tan \beta_o l_1} - \cot \beta_o l_2 \right] \quad (3.43)$$

where β_o is the propagation constant at the odd-mode resonance angular frequency ω_o .

Using equation (3.38) and equation (3.43), it leads to

$$\frac{2C_m \omega_o}{Y_c} = \frac{1}{\tan \beta_o (l_1 + l_2)} \quad (3.44)$$

For a small C_m , $(l_1 + l_2) \approx \lambda_g / 4$, the right hand side of equation (3.44) can be expanded by using Taylor expansion at $(l_1 + l_2) = \lambda_g / 4$. By omitting the high-order terms, equation (3.44) can be rewritten as

$$\frac{2C_m \omega_o}{Y_c} + \beta_o (l_1 + l_2) = \frac{\pi}{2} \quad (3.45)$$

Since

$$\beta_o = \frac{\omega_o \sqrt{\epsilon_{re}}}{c}$$

the odd-mode resonant angular frequency can be determined by

$$\omega_o = \frac{\pi}{2(2C_m Z_c + A)} \quad (3.46)$$

The center frequency of the BPF can be approximated by averaging the even- and odd-mode frequencies as

$$f_0 = \frac{1}{4\pi} (\omega_e + \omega_o) = \frac{1}{8} \left[\frac{1}{2C_m Z_c + A} + \frac{1}{2Y_c L_m + A} \right] \quad (3.47)$$

The coupling between two modes is characterized by the coupling coefficient C [7] which can be computed from the knowledge of even- and odd-mode frequencies as

$$C = \frac{\omega_o^2 - \omega_e^2}{\omega_o^2 + \omega_e^2} = F(Y_c L_m - C_m Z_c) = M - E \quad (3.48)$$

where

$$F = \frac{4(A + Y_c L_m + Z_c C_m)}{(2Y_c L_m + A)^2 + (2Z_c C_m + A)^2},$$

$$M = F Y_c L_m \text{ and } E = F Z_c C_m$$

From equation (3.48), the inter-stage couplings of the SEMCP are formed by two separated parts: electric (E) coupling and magnetic (M) coupling as illustrated in filter topology of Figure 3.40 b). The couplings in the two coupling paths are dependent on each other and have the canceling effects for the total coupling coefficient C . Thus different coupling coefficient can be realized by changing the values of two coupling elements. The external quality factor [23] can be determined by averaging the external quality factor of the even- and odd- modes:

$$Q_e = \frac{1}{2}(Q_{e,e} + Q_{e,o}) = \frac{\pi Z_L}{4 Z_c} \left(\frac{1}{\sin^2(\beta_e(l_2 + l_e))} + \frac{1}{\sin^2 \beta_o l_2} \right) \quad (3.49)$$

where

$$l_e = \frac{2Y_c L_m c}{\sqrt{\epsilon_{re}}}$$

$Z_L = 1/Y_L$ is the impedance at the I/O ports.

The transmission characteristic of the filter circuit can be calculated from the odd- and even-mode input admittances and is expressed as

$$S_{21} = \frac{\bar{Y}_{odd} - \bar{Y}_{even}}{(1 + \bar{Y}_{odd})(1 + \bar{Y}_{even})} \quad (3.50)$$

where $\bar{Y}_{even} = Y_{even}/Y_L$ and $\bar{Y}_{odd} = Y_{odd}/Y_L$ are normalized even- and odd-mode admittances respectively. Matlab 6.5 is used to discuss two useful cases:

i) Magnetic coupling M dominant ($C > 0$)

Now $Y_c L_m > C_m Z_c$, the operating frequency of the odd-mode is higher than that of the even-mode and the coupling in the lower coupling path of Figure 3.40 a) is dominant. The weak electric coupling can affect both the bandwidth and the characteristics of the stopband. For the same dimension of the transmission line and I/O ports, the effects of coupling elements (L_m and C_m) in the two coupling paths on filter performance are compared in Figure 3.43. When $C_m=0$, there is only one zero point generated in the upper stopband due to harmonic effects. While $C_m>0$, there is an additional zero point generated in high stopband and the filter demonstrates a good skirt selectivity in the upper stopband. When both L_m and C_m are increased, the operating frequency is shifted downwards while keeping the coupling coefficient. However, the rejection in the stopband will be reduced. When L_m is fixed at $L_m=0.15\text{nH}$, increasing C_m (see $C_m=0\text{pF}$ and $C_m=0.01\text{pF}$) can decrease the bandwidth of the filter by reducing the operating frequency of the odd-mode (refer to Figure 3.43 and equations (3.46) and (3.48)).

Figure 3.43 The spectrum responses of M dominant SEMCP filters

ii) **Electric coupling E dominant ($C > 0$)**

Now $Y_c L_m < C_m Z_c$, the operating frequency of the even-mode is higher than that of the odd-mode and the coupling in the upper coupling path of Figure 3.40 a) is dominant. The weak magnetic coupling path can also affect both the bandwidth and the characteristics of the stopband. For the same dimension of the transmission line and I/O ports, the effects of coupling elements (L_m and C_m) in the two coupling paths on filter performance are compared in Figure 3.44. When $L_m=0$, no finite zero point can be generated in the stopband. When $L_m>0$, there are two additional zero points generated in stopband and the filter demonstrates a steep skirt selectivity. When both L_m and C_m are increased, the operating frequency is shifted downwards, while the coupling coefficient is kept constant. However, the rejection in the stopband will be poorer. When C_m is fixed at 0.05pF, increasing L_m (see $L_m=0.0$ nH and $L_m=0.05$ nH) decreases the bandwidth of the filter by reducing coupling coefficient C (refer to Figure 3.45 and equations (3.42) and (3.48)).

Figure 3.44 The spectrum responses of E dominant SEMCP filters

By comparing Figure 3.43 with Figure 3.44, it is interesting to note that for the SEMCP filter with the same transmission line dimensions, the filter response can be completely controlled by choosing different dominant coupling path as well as different coupling element values. The canceling effect of couplings in two different coupling paths fully exploited for designing a filter with narrow bandwidth, where a smaller coupling coefficient is required. It is important to have an additional zero point generated in either the low stopband or the upper stopband. Thus the frequency responses of the quarter-wavelength filter become more controllable. The higher stopband rejection can be achieved by using smaller values of the lumped elements. These advantages will be helpful in the filter designs as demonstrated in following sections.

3.6.2 Second-order OG filter realization

Figure 3.45 The configurations of the second-order SEMCP filters

For most general applications, the lumped capacitor can be implemented by a

coupled gap or coupled transmission lines. The lumped inductance can be realized by a via hole [12]-[13] or a section of high characteristic impedance transmission line connected with ground via hole in series (See four perturbations in section 3.2). As in Figure 3.45, the three configurations of the second-order SEMCP filters with tap feeders are illustrated. The filter topology is shown in Figure 3.40 b). Figure 3.45 i) illustrates the condition when there are a weak electric coupling, generated by the open-ended coupling gap, and a magnetic coupling, generated by the common via ground, in inter-stage coupling. Figure 3.45 ii) shows the case when both electric coupling, generated by parallel coupling lines with the length of l_J , and magnetic coupling, generated by a section of high characteristic impedance transmission line (with length of l_h and width of w_h) connected with ground via hole in series, are increased as compared to Figure 3.45 i). By properly choosing the dimensions of the coupling parts, either magnetic or electric coupling can be made dominant. Figure 3.45 iii) presents the configuration where parallel coupling lines and via ground are used. The RT/Duroid 6010 dielectric substrate with relative dielectric constant of $\epsilon_r = 10.2$ and thickness of 0.635mm is used for the design. Commercial EM software IE3D is used for the analysis. Figure 3.46 demonstrates the condition when the electric coupling is dominant (i.e. coupling coefficient C is smaller than zero). The length of l_u , which mainly determines the occupied area of the filter, is fixed. The filters with two coupling paths can generate an additional zero point in the rejection band, which is similar to the lumped elements condition specified in Section 3.6.1. When the external quality factor is fixed, the increased magnetic and electric

couplings cancel each other in the operating frequency range. Therefore, the filter bandwidth, which is mainly determined by the inter-stage coupling coefficient C , changes very little, while the operating frequency is shifted downwards due to the increased loadings in two coupling paths. Figure 3.47 gives the condition when magnetic coupling is dominant (i.e. coupling coefficient C is larger than zero). When the length of each resonator is fixed, the filter of SEMCP generates an additional ZP in the upper rejection band. The increased E and M couplings also cancel each other. Thus the filter bandwidth is maintained even the electric and magnetic couplings are increased. However, the operating frequency is shifted downwards due to increased coupling in each path because of the loading effects in two coupling paths.

Figure 3.46 The frequency responses of the second-order E dominant SEMCP filters

a) configuration of Figure 3.45 ii) $l_b=1.5$ mm, $l_u=8.1$ mm; $l_f= 6.2$ mm $R_{via}=0.1$ mm $S=0.1$ mm, $l_h=0.4$ mm, $w_h=0.2$ mm b) configuration of Figure 3.45 iii) $l_b=1.5$ mm, $l_u=8.1$ mm; $l_f= 3.2$ mm $R_{via}=0.1$ mm $S=0.1$ mm c) configuration of Figure 3.45 iii) $l_b=1.5$ mm, $l_u=8.1$ mm; $l_f= 1.7$ mm $R_{via}=0.3$ mm $S=0.1$ mm

Figure 3.47 The frequency responses of the second-order M dominant SEMCP filters

a) configuration of Figure 3.45 ii) $l_b=1.5$ mm, $l_u=11.9$ mm; $l_j= 1.15$ mm $R_{via}=0.1$ mm $S=0.1$ mm, $l_h=0.4$ mm, $w_h=0.2$ mm b) configuration of Figure 3.45 iii) $l_b=1.5$ mm, $l_u=11.9$ mm; $l_j= 0.5$ mm $R_{via}=0.1$ mm $S=0.1$ mm c) configuration of Figure 3.45 i) $l_b=1.5$ mm, $l_u=12.3$ mm; $G= 0.1$ mm $R_{via}=0.1$ mm $S=0.1$ mm

In Figure 3.48, both the simulation and measurement results of a magnetic coupling dominant filter in Figure 3.45 i) are compared. For the magnetic coupling dominant condition, there are two finite transmission ZPs generated in the upper stopband. The first zero point, close to the operating frequency, is generated by the canceling effect of the electric and magnetic coupling. The second ZP is mainly due to the harmonic effects of the distributed transmission line. For the narrow bandwidth BPF, the first ZP is located quite close to the operating frequency, thus the skirt selectivity at the upper stopband can be greatly improved by the finite transmission ZP. The measured specifications of the filter are: the operating frequency is 1.94 GHz, -3 dB bandwidth is 160MHz, insertion loss is 1.75 dB and stopband rejection is larger than 30 dB from 2.12 GHz to 2.56 GHz. The rejections at the first zero point 2.15GHz

and the second zero point of 2.4 GHz are 33 dB and 47 dB respectively. The fabricated photographs of the filter with reference to the ball pen are illustrated in Figure 3.49.

Figure 3.48 Simulated and measured results of second-order M dominant filter

Figure 3.49 The photograph of the fabricated M dominant SEMCP filter

In Figure 3.50, the results of simulation and measurement for electric coupling dominant filter in Figure 3.45 iii) are compared. The extracted inductance value of via model in IE3D is smaller than that from HFSS and measurement (refer to discussion in Section 3.4.3). Hence, larger bandwidth and lower attenuation in the lower

stopband can be seen in the measured results. The fabrication errors are also possibly generated in the single via hole, which is realized by vertically drilling and electroplating. The fabrication accuracy of the via is more difficult to be controlled than that of the planar metal trace or the gap. The measured specifications of the filter are: the operating frequency is 2.4 GHz, 1 dB bandwidth is 380MHz and the insertion loss is 0.85dB. The occupied area is only $0.065\lambda_0 \times 0.027\lambda_0$ (λ_0 is the free space wavelength at operating frequency). Figure 3.51 shows the fabricated filter.

Figure 3.50 Simulated and measured results of the second-order *E* dominant filter

Figure 3.51 Photograph of the *E* dominant second-order SEMCP filter

Size: $0.81 \times 0.34 \text{ cm}^2$ (0.275 cm^2)

3.6.3 General higher order SEMCP OG filter

Figure 3.52 The topology of the high-order SEMCP filter

The SEMCP configuration can also be used to design higher order BPF with cascaded resonators. The equivalent topology of an n -stage SEMCP filter is shown in Figure 3.52. In the filter topology, the SEMCPs exist in interval, and the inter-stage coupling coefficient in SEMCP structures with complicated distributed E and M couplings can be determined by using the full-wave EM software. Procedure in [8] is used to determine the parameters. The coupling parameters can be calculated by

$$C = \frac{\omega_{high}^2 - \omega_{low}^2}{\omega_{high}^2 + \omega_{low}^2} \quad (3.51)$$

where ω_{high} and ω_{low} are higher and lower resonant frequencies of two coupled resonators respectively.

As introduced in Section 3.6.1 and 3.6.2, both electric and magnetic coupling could be dominant. The dominant electric coupling can be set in order to generate zero points in the lower rejection band so that the skirt selectivity and the rejection can be improved. On the other hand, the dominant magnetic coupling can be used to improve the skirt selectivity and the rejection in the upper rejection band. If the filter has both dominant E and M SEMCPs simultaneously, the skirt selectivity in both sides near to

the passband can be improved and multiple ZPs will be generated in both sides of the stopband. The filter shows the quasi-elliptical response in the spectrum.

Figure 3.53 The configuration and topology of the fourth-order SEMCP filter

The configuration and the topology of a fourth-order SEMCP BPF are shown in Figures 3.53 a) and b) respectively. The SEMCPs are introduced between resonators 1 and 2 as well as 3 and 4. To achieve compact size and sharp skirt selectivity in the lower stopband, the dominant electric couplings are chosen in the fourth-order SEMCP filter. Firstly, a filter with 17% fractional bandwidth (FBW) is designed. The external quality factor Q_e and the inter-stage coupling C can be calculated from equation (3.54) and (3.56) respectively. $Q_e = 6.52$ and

$$\bar{C} = \begin{bmatrix} 0 & -0.141 & 0 & 0 \\ -0.141 & 0 & 0.112 & 0 \\ 0 & 0.112 & 0 & -0.141 \\ 0 & 0 & -0.141 & 0 \end{bmatrix}$$

where the negative signs of coupling coefficients in the coupling matrix $\bar{C}$ denote the electric coupling as in equation (3.53). The optimized dimensions of the filter are: $W_0=0.59$ mm, $l_b=3.95$ mm, $l_u=5.94$ mm, $l_f=4.52$ mm, $R_{via}=0.2$ mm, $S=0.09$ mm, $S_2=0.6$ mm. To demonstrate the control of the ZPs on the lower and higher stopbands, the configurations and EM simulation results of the fourth-order SEMCP filters with one E dominant pair and one M dominant pair are shown in Figure 3.54. Both the inverted and no-inverted SEMCP pairs have the same dimensions. Each filter generates two additional ZPs in the lower and upper stopbands as compared to the traditional inter-digital filter. One more ZP is generated in the low stopband of on-inverted SEMCP pairs. The configuration of case a) in Figure 3.54 is a mixed coupling path between the E and M dominant pair (because the pairs are inverted), while the configuration in case b) of Figure 3.54 is actually a separated E dominant mixed coupling path between the E and M dominant pair (or in the sequence of E - E - M dominant because the pairs are not inverted). Thus there are two ZPs generated in the lower stopband and one ZP generated in the upper stopband.

From the above cases, it can be seen that our general coupling topology is quite useful for the cascaded resonators to generate multiple ZPs to obtain the elliptical response without using the canonical cross coupling topology among the resonators [15]-[18], [21]. It should be useful to control the stopband of the filter and can be employed as an alternative way to the traditional cross coupling techniques.

Figure 3.54 The configuration and topology of the fourth-order SEMCP filter

Dimension (For both structure a) and b)): $W_0=0.59$ mm, $l_b=2.9$ mm, $l_u=5.94$ mm, $l_f=4.52$ mm $R_{via}=0.1$ mm $S_1=0.3$ mm, $S_2=0.9$ mm, $l_h=1.2$ mm, $w_h=0.4$ mm

The measured results and the simulated results of the filter in the frequency range of 500 MHz~3 GHz are compared in Figure 3.55. The operating frequency (3.18 GHz) of the measured results is about 100MHz lower than that of the simulated results (3.29 GHz). The measured bandwidth BW=430 MHz (FBW=19.7%) is larger than simulated bandwidth 390 MHz (FBW=17%). The measured minimum insertion loss is 1.4 dB, and the return loss is better than 12 dB in 1.98 GHz~2.33 GHz. The rejection at 1.97 GHz (−3 dB edge of the filter in low side band) is −3 dB, while the rejection at 1.7 GHz (0.63BW) is larger than −40 dB. Figure 3.56 shows the simulated results and the measured results in the frequency range DC~7.05 GHz. There are two zero points (1.77 GHz with rejection of 53.8 dB and 1.11 GHz with rejection of 73.2 dB) in the lower rejection band of the simulated results. Both simulated and measured

results show that the filter has good skirt selectivity in the lower rejection band. The discrepancy of the operating frequency may come from the via holes as explained before, which have effect on both the bandwidth and the frequency of the zero point.

Figure 3.55 Simulated and measured results of the fourth-order *E* dominant filter

To compare the SEMCP filter with the traditional hairpin-line filter, a fourth-order hairpin-line filter with the bandwidth of 760 MHz (BW=34%) has also been designed and fabricated. The measured results of the fourth-order SEMCP filter and the fourth-order hairpin-line filter are compared in Figure 3.56. The measured minimum insertion losses for SEMCP and hairpin-line filters are 1.4 dB and 1.48 dB respectively. The SEMCP filter has narrower bandwidth, wider stopband bandwidth, and better skirt selectivity in the lower rejection band. To reduce the bandwidth of the hairpin-line filter, the space between the adjacent resonators should be further apart and this may lead to increase of filter size [29]. It should be addressed that the higher order topology of SEMCP is different from the general filter topology in [30]-[35].

Because there are two separated E and M coupling paths exist between adjacent resonators in the SEMCP filter topology, while the general filter coupling topology in the referenced literature has only one coupling path between any two adjacent resonators. It can be seen that the SEMCP has advantages in designing filter with low insertion loss and compact size. The photograph of the SEMCP filter and hairpin-line filter are illustrated in Figure 3.57. The size of the hairpin line filter (without considering the I/O feed lines, $1.26 \text{ cm} \times 1.48 \text{ cm} = 1.865 \text{ cm}^2$) is three times larger than that of the SEMCP filter $0.06\lambda_0 \times 0.0545\lambda_0$ ($0.82 \text{ cm} \times 0.75 \text{ cm} = 0.615 \text{ cm}^2$). The group delays of the fourth-order SEMCP filter and hairpin-line filter are compared in Fig.3.58. Because the rolloff improvement of the filter is at expense of its worse flatness of group delay, the group delay flatness of SEMCP filter is worse than that of the traditional comb filter. Fortunately, the group delay variation of $\pm 1 \text{ ns}$ is better than the requirement of $\pm 5 \text{ ns}$ in most of high data rate digital communication system.

Figure 3.56 Measured results of the fourth-order SEMCP filter and hairpin filter

Figure 3.57 The photograph of the fabricated fourth-order SEMCP and hairpin filters

Figure 3.58 Group delays of the measured fourth-order SEMCP and hairpin filters

3.7 Conclusion

In this chapter, the MS OG structures are extensively investigated. The coupling characteristics, ZP generation and spurious suppression, which are important for the filter design, are analyzed. The methodologies for additional ZP generation, size reduction and stopband extension are also discussed and demonstrated. The model-based analysis, transmission line network and the odd- and even- mode analyses are employed in the filter analysis and design. Novel filter and diplexer are

implemented using the traditional PCB technology. A novel general SEMCP filter topology together with the configuration is proposed. The characteristics of the E dominant and M dominant SEMCP filters are investigated and the advantages such as compact size, sharp skirt selectivity and low loss are demonstrated by both simulation and experiment. The general higher order SEMCP filter topology is introduced and the fourth-order E dominant filter with compact size and low insertion loss is implemented. It should be noted that though the topology is realized by using microstrip line, it can be applicable to other transmission lines such as the coplanar waveguide (CPW), stripline. For example, if the CPW transmission line is used to realize the filter topology, the grounding via hole can be replaced by ground shunt. The general high-order SEMCP topology and configuration are useful techniques to control the stopband response of the filter. It should be an alternative method to cross coupling techniques to generate elliptic response using cascaded resonators.

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CHAPTER 4

PERIODIC LOADING LINES AND FILTERS USING THIN-FILM TECHNOLOGY

As discussed in the review of Chapter 2, the periodic guiding structures [1]-[9] have drawn much attention to microwave and optical applications. Because of four important aspects: 1) periodic structures lead to the distinctive electromagnetic propagations in both the passband and the stopband [1]-[2]; 2) periodic structures can support the so-called slow-wave which can be used to reduce the circuit size [4]-[5]; 3) periodic structures can be used to design the electromagnetic bandgap (EBG) or photonic bandgap (PBG) circuits. By using this characteristic, some passive circuits, such as leaky wave antennas, frequency selected surface (FSS) and filters [5]-[7] can achieve very good performance; 4) periodic structures are easier to be analyzed and modeled than non-periodic ones [8]-[9].

Figure 4.1 Structure of conductor-backed coplanar waveguide (CBCPW)

■ Metal ■ Substrate

Filters based on coplanar waveguide (CPW) [5], [7] have been proven to be more suitable than MS filters for monolithic microwave integrated circuits (MMICs) [2]. The conductor-backed CPW (CBCPW), as illustrated in Figure 4.1, has the

advantages, such as better heat sinking ability, better mechanical strength, and lower characteristic impedance etc., over the conventional CPWs [2]. Therefore it is attractive to be used in MMICs and the circuit packaging [10]-[11]. However, there are two aspects that limit the utilization of the CBCPW in the filter implementation. Firstly, the filter size is relative large below millimeter wave frequencies [11]-[12], and the CBCPW filter can not be economically designed, especially for MMICs. Secondly, the CBCPW is known to support unwanted bulk modes [10]-[17]. These bulk modes can deteriorate the circuit performance. Much work has been done on the guiding wave characteristics of the CBCPW and its modified structures [12]-[17]. However, most of the work in the literature had been focused on the transmission characteristics, shielding effects and radiation of the CBCPW transmission lines. Little work has been done on the gradually periodic loading on CBCPW transmission lines and the filters.

In this chapter, EM-based cell parameter extracting and cascading cell with phase compensation are introduced to investigate and design periodically loaded CBCPW lines and BPFs with compact size. EM-based cell parameter extracting is good at extracting the distributed parameters of the periodic cell with gradual change. The cascading cell with phase compensation can be used to design the lines and filters with low computation and acceptable accuracy. It is also demonstrated that the “finite ground” [2], [10]-[11] of the CBCPW lines with loading or without loading can be used to achieve the leakage-mode-free transmission lines in a wide frequency range. However, “finite ground” for leakage-mode-free is not applicable to the complicated

resonant circuits such as the end-coupled BPFs. It is shown by our experimental results that the proper interval of the ground via holes is an effective way to suppress the power leakage in the CBCPW BPFs.

4.1 Parameter extracting of the Unit Cells

A section of CBCPW transmission line or its discontinuities (called unit cell), whose physical length is equal to or smaller than one-eighth of the guided wavelength at the operating frequency, is most commonly modeled as one or several lumped elements approximately. These equivalent lumped elements can be used to characterize transmission lines or their discontinuities for understanding of the operating mechanism and realizing the circuits. However, for the gradual changes of the loadings or the discontinuities, the analysis of capacitively loading using the lumped-circuit model in Chapter 2 is less accurate due to the distribution effects. Furthermore, the leakage characteristics are unpredictable. Therefore, the full-wave EM extracting methods should be adopted to get the frequency-dependent equivalent element values. The unit cells, classified as the transmission cells and the coupling cells in the following parts, have the same physical length of l_i . The lateral ground before being defected is set to the same finite width of W_g (please refer to Figure 4.1). For the transmission line and the filter with low loss, the phase delay and the resonant frequency with and without consideration of the smaller loss show no significant difference. Therefore, the phase delay and the resonant frequency characteristics are mostly concerned [18], [19]. Thus only the equivalent reactance of the equivalent

lumped circuit model is extracted from the cells in the following analysis. The IE3D software is used in EM simulation. The Asfired Alumina substrate ($\epsilon_r=9.8$, $H=0.38$ mm, $t=4$ μ m) and the thin-film fabrication technology from Applied Thin-film Products (ATP) company are used in the designs.

4.1.1 Parameter extracting of the transmission cells

Figure 4.2 Transmission cells and their lumped models

- a) Unloaded cell b) Loaded inductive transmission (IT) cell
c) Loaded capacitive transmission (CT) cell d) Equivalent circuit for a), b) and c)

Unit transmission cells (T-cells) are demonstrated in Figure 4.2. The cell length of l_i is denoted as the physical length between two de-embedding reference planes T and T' . R_o is the radius of the circular defect, centered at the middle of the cell (l_i , W_0). R_i is the radius of the circular patch with the same center as the circular defect. 50 Ω system is chosen as the characteristic impedance of the CBCPW transmission line at the input and output ports. The center conductor's width W_0 is 0.1mm and slot width S_0 is 0.05mm. The unloaded unit cell is shown in Figure 4.2 a). The structures

shown in Figure 4.2 b) & c) can be considered as the unit cells with gradually loading (loaded) transmission lines. The characteristics of unit cells can be extracted by de-embedding the I/O ports with respect to the reference plane T and T'.

Figure 4.3 Phase characteristics of the three kinds of cells

When $R_o=0.26\text{mm}$, $R_i=0.21\text{mm}$, and $l_i=0.46\text{mm}$, the characteristic impedance of the loaded CT cell varies from 50Ω to 31.4Ω and then varies from 31.4Ω back to 50Ω . For the loaded IT cell, the characteristic impedance gradually changes from 50Ω to 72.5Ω and then gradually changes back to 50Ω . The phase delay characteristics of the three types of unit cells with the same length of l_i are compared in Figure 4.3 by using IE3D software. The loaded cells generate larger phase delay than the unloaded one. The CT cell has the largest phase delay among the three cells.

These T-cells can be modeled as a π -network as shown in Figure 4.2 d) with different element values. The S-parameters are calculated by using full-wave EM

IE3D software, and then Y-parameters of the equivalent π -network can be calculated from the S-parameters as [21]

$$\frac{Y_b}{Y_0} = \frac{2S_{21}}{(1 + S_{11})(1 + S_{22}) - S_{21}S_{12}} \quad (4.1)$$

$$\frac{Y_a}{Y_0} = \frac{(1 - S_{11})(1 + S_{22}) + S_{21}S_{12} - 2S_{12}}{(1 + S_{11})(1 + S_{22}) - S_{21}S_{12}} \quad (4.2)$$

Because the two ports are reciprocal, equations (4.1) and (4.2) can be rewritten as

$$\frac{Y_b}{Y_0} = \frac{2S_{21}}{(1 + S_{11})^2 - S_{21}^2} \quad (4.3)$$

$$\frac{Y_a}{Y_0} = \frac{1 - S_{11} - S_{21}}{1 + S_{11} + S_{21}} \quad (4.4)$$

The imaginary parts (refer to Figure 4.2) of the extracted Y-parameters are used to extract the equivalent lumped element values as shown in Figure 4.4.

Figure 4.4 Frequency-dependent element values of π -network models of T-cell

As compared with the unloaded cell, the IT cell has a larger equivalent serial inductance and shunt capacitance, and this is contrary to the case of the CT cell. At

the low operating frequency range (below 8GHz), the values of the equivalent circuit elements almost remain constant with respect to the frequency. As the frequency increases, the frequency dispersion effects also increase due to the fringe field and increased electrical size as compared to the wavelength. Thus the values of the equivalent element change as the demonstrated in Figure 4.4.

4.1.2 Parameter extracting of the coupling cells

Unit coupling cells (C-cells) are illustrated in Figure 4.5. The cell length of l_i is denoted as the physical length between two de-embedding reference planes T and T' . The difference of the C-cells from the T-cells is that the gap or shunt exists in the middle of the coupling cells. The gap or shunt can be treated as an inverter (refer to Chapter 2). The application of these cells will be introduced in the following. The structures of the unloaded and loaded capacitive coupling (CC) cells are shown in Figures 4.5 a) and b) respectively. The structures of the unloaded and loaded inductive coupling cells (IC) are illustrated in Figures 4.5 c) and d) respectively. The coupling coefficients of the structures in Figures 4.5 a) and c) are limited by the minimum widths of the gap and the shunt, which can be manufactured by the fabrication technology, especially for the miniaturized transmission line where the center trace of W_0 is quite narrow. The loaded IC cell and CC cell have larger coupling coefficients as compared to the unloaded counterparts and therefore can more easily satisfy the stringent requirements of the fabrication. The length of the cell between two de-embedding reference plane T and T' is still $l_i=0.46$ mm. S_l and W_s denote the gap

width and the shunt width respectively. The loaded and unloaded CC cells can be modeled as π -networks as shown in Figure 4.5 e), and the loaded and unloaded IC cells can be modeled as T-networks as shown in Figure 4.5 f).

Figure 4.5 Coupling cells and the lumped models:

- a) Unloaded capacitive coupling (CC) cell
- b) Loaded capacitive coupling (CC) cell
- c) Unloaded inductive coupling (IC) cell
- d) Loaded inductive coupling (IC) cell
- e) Equivalent circuit of a) or b)
- f) Equivalent circuit of c) or d)

The similar de-embedding procedure is used to obtain the S-parameters of the unit coupling cells. The element values of the equivalent π -network of the CC cells are extracted by using equations (4.3) and (4.4). The T-network circuits corresponding to the IC cell are extracted by using equations (4.5) and (4.6) as [21], given by

$$\frac{Z_b}{Z_0} = \frac{2S_{21}}{(1 - S_{11})(1 - S_{22}) - S_{21}S_{12}} \quad (4.5)$$

$$\frac{Z_a}{Z_0} = \frac{(1 + S_{11})(1 - S_{22}) + S_{21}S_{12} - 2S_{21}}{(1 - S_{11})(1 - S_{22}) - S_{21}S_{12}} \quad (4.6)$$

Since two ports are also symmetrical, equations (4.5) and (4.6) can be rewritten as

$$\frac{Z_b}{Z_0} = \frac{2S_{21}}{(1 - S_{11})^2 - S_{21}^2} \quad (4.7)$$

$$\frac{Z_a}{Z_0} = \frac{1 + S_{11} - S_{21}}{1 - S_{11} + S_{21}} \quad (4.8)$$

Figure 4.6 Frequency-dependent lumped element values of CC cell models

The extracted parameters (refer to Figure 4.5) for $S_l=0.02$ mm are shown in Figure 4.6. The equivalent element values C_a and C_b of the loaded CC cell, which has the same coupling gap S_l as the unloaded one, are compared with those of the unloaded CC cell. The C_b of the loaded CC cell is four times as large as the C_b of the unloaded CC cell, while the C_a of the loaded CC is about two times as large as the C_a of the unloaded CC cell. The equivalent capacitances C_a and C_b of the loaded cell increase with the increase of the frequency. However the C_a and the C_b of the

unloaded cell remain constant. It is because that for both the unloaded CC cell and the loaded CC cell, the frequency dispersion effects increase with the increase of the frequency, while the frequency dispersion effects of the loaded CC cell are larger than that of the unloaded CC cell. Figure 4.7 demonstrates the extracted element values of the IC cells when $W_s=0.14\text{mm}$. The element values of the extracted inductances L_a and L_b of the loaded IC cell are much larger than those of the unloaded one.

Figure 4.7 Frequency-dependent lumped element values of IC cell models

Both the loaded and the unloaded cells (i.e. IC cell and CC cell) have the same gap size S_l and shunt width W_s . The large coupling element values imply that for the same coupling coefficient, the loaded cells have larger gap S_l for the CC cell and larger width W_s for the IC cell. Therefore, the loading at the coupling parts can more easily satisfy the stringent requirement of the minimum width of the gap and trace for fabrication processes especially, for the end-coupled filter.

At the same time, the loaded CC and IC cells, which are used as the inverters for

the end-coupled filters as in Chapter 2, also contribute to the size reduction due to the increased equivalent electric lengths of pha for the coupling structures in the J and K inverters. The K inverter corresponding to Figure 4.5 f) can be written as [20]

$$K = Z_0 \left| \tan \left(\frac{pha}{2} + \tan^{-1} \frac{Z_{ai}}{Z_0} \right) \right| \quad \text{ohms} \quad (4.9)$$

$$pha = -\tan^{-1} \left(\frac{2Z_{bi}}{Z_0} + \frac{Z_{ai}}{Z_0} \right) - \tan^{-1} \frac{Z_{ai}}{Z_0} \quad \text{radians} \quad (4.10)$$

The calculated normalized K and pha are illustrated in Figure 4.8. The pha of the loaded IC cell is one and a half times larger than that of the unloaded IC cell. The K value of the loaded IC cell is about four times that of the unloaded IC cell.

Figure 4.8 Normalized K -inverter value and equivalent electric length of IC cell

Note: f_0 is the operating frequency in GHz

The J -inverter corresponding to Figure 4.5 e) can be written as [20]

$$J = Y_0 \left| \tan \left(\frac{pha}{2} + \tan^{-1} \frac{Y_{ai}}{Y_0} \right) \right| \quad \text{mhos} \quad (4.11)$$

$$pha = -\tan^{-1} \left(\frac{2Y_{bi}}{Y_0} + \frac{Y_{ai}}{Y_0} \right) - \tan^{-1} \frac{Y_{ai}}{Y_0} \quad \text{radians} \quad (4.12)$$

The calculated normalized J and pha are presented in Figure 4.9. The pha of the loaded CC cell is about two times as large as that of the unloaded CC cell. The J factor for the loaded CC cell is about four times as large as that of the unloaded CC cell.

Figure 4.9 Normalized J -inverter value and equivalent electric length of CC cell

Note: f_0 is the operating frequency in GHz

From the above comparison between the loaded and unloaded cells, it is clear that the equivalent element values of the loaded cells are different from those of the unloaded cells. The loaded T cells have larger phase delay than the unloaded T cells. For the coupling cells, the loading contributes to both coupling enhancement by increasing the absolute values of K and J inverter and size reduction by increasing the absolute value of the equivalent electrical length of pha of the K and J inverters. The extracted frequency-dependent K and J inverters are different from constant K and J inverters in Chapter 2. The loaded cells can be used to reduce the filter size and increase the inter-stage coupling as to be demonstrated in Section 4.3.

4.2 Phase Compensation of CBCPW Lines

Figure 4.10 Top metal patterns of the 9-cells CBCPW transmission lines

■ Substrate ■ Metal a) Unloaded b) IT cell line c) CT cell line

The loaded and unloaded CBCPW transmission lines are shown in Figure 4.10. All these transmission lines have the same total physical length. Each of them is formed by nine cells with the same unit length of l_i and very short connection lines in the I/O ports for the touching of the probe-tips in wafer-based measurement as shown in Figure 4.10. The structure shown in Figure 4.10 a) is the unloaded CBCPW transmission line with the side ground width of W_g , the structure shown in Figure 4.10 b) is an IT cell line, and the structure shown in Figure 4.10 c) is a CT cell line. $Z_0=1/Y_0= 50\Omega$ is chosen as the characteristic impedance of the CBCPW transmission line at the input and output ports.

The measured transmission, reflection and phase delay characteristics of the

nine-cell CBCPW lines with and without loading, as shown in Figure 4.10, are illustrated in Figure 4.11, Figure 4.12 and Figure 4.13, respectively. In Figure 4.11, the insertion losses of the three types of the cell lines are compared. For the operating frequency from 500MHz to 1.43GHz, the CT cell line has the lowest insertion loss among the three lines. If the operating frequency is below 500MHz, the insertion losses of the unloaded and the IT loaded cell lines are almost the same. The reason is that, at lower operating frequencies, where the metal loss is dominant, CBCPW line with a narrow width of the center strip has larger metal resistance in DC and low frequency range than the CBCPW line with a wide center strip. As the frequency increases beyond 1.43GHz, the CT and IT cell lines generate more losses than the unloaded line and there are ripples generated in the transmission response. The reason is that the discontinuity of those of the loaded lines leads to more reflection and radiation as demonstrated in Figure 4.12. From Figure 4.13, it can be observed that both the CT cell line and the IT cell line have larger phase delay than the unloaded one. To achieve both low radiation loss and large phase delay, the optimization of the loaded cell with the consideration of the matching is needed in certain frequency ranges.

Figure 4.11 Measured transmission characteristics of periodic cell lines

Figure 4.12 Measured reflection characteristics of periodic cell lines

Figure 4.13 Measured phase delay of periodic cell lines

4.2.1 Discussion on leakage modes

Assuming transmission lines formed by p number of cascading T-cells (the structures in Figure 4.10 are the special cases of $p=9$). As shown in Figure 4.14, when the thin substrate with the thickness of H is used, the ground metal at the back side of the CBCPW is introduced in close proximity to the CPW field. The back side ground metal can effectively change the propagating behaviors of the EM wave, and the unwanted bulk modes (including the surface, parallel plate (PP) and MS-like modes) will be excited [12]-[16]. For the simple case of a rectangular patch as shown in Figure 4.10 a), the resonant frequency formula in [13] can be modified to predict the leakage modes as

Figure 4.14 Substrate reflections and radiations of PP mode in finite CBCPW

$$f_{mn} = \frac{c}{2\pi\sqrt{\epsilon_r}} \sqrt{\left(\frac{n\pi}{pl_i}\right)^2 + \left(\frac{m\pi}{W_g}\right)^2} \quad (4.13)$$

where c is the velocity of light, ϵ_r is the relative dielectric constant, W_g is the width of the top ground plane, l_i is the length of each unit cell, and m, n denote the mode number.

From equation (4.13), a small W_g and pl_i can result in a large frequency f_{mn} of the leakage mode. If f_{mn} is larger than the operating frequency of the cell line, the cell

line can yield the leakage-mode-free operation. Properly choosing W_g can make the “finite ground” CBCPW exhibit no resonance in the wider frequency range as explained in [13]. For our interested frequency range of DC to 30 GHz, no PP mode ($m \geq 0.5$) is excited from the unit cell to the cascaded ten cells as illustrated in Figure 4.15. When the cell number p is larger than 3, the MS mode and CPM-mode [11], which have the operating frequencies corresponding to $m=0$ in equation (4.13), may exist. However, the electromagnetic power coupled to the CPM-mode and microstrip mode can be neglected if the characteristic impedance of these modes in the “finite ground” CBCPW is much higher or much lower than 50Ω [16]. This is helpful to explain the leakage-mode-free in these loaded and unloaded transmission lines with the finite lateral grounds.

Figure 4.15 Relationship of the cell number with possible resonant frequency

The measured results of the nine-cell transmission lines with and without ground

vias are compared in Figure 4.16 and the practical photograph is given in Figure 4.10. From DC to 30.05 GHz, the insertion losses of these loaded and unloaded transmission lines are less than 1 dB. It is clear that the CBCPW cell lines with via (called “via ground”) and without via (called “finite ground”) are leakage-mode-free in the investigated frequency range. However, for the complicated CBCPW structures such as the end-coupled filters, which are operating in resonant condition, the “finite ground” of CBCPW becomes impractical as to be discussed in the Section 4.3.2.

Figure 4.16 Measured results of the lines in Figure 4.10 with and without ground via

4.2.2 Phase compensation method

The S-parameters calculated by the traditional cascading network parameters of cells actually neglect the space and substrate coupling among the cells. The cascading errors will be generated by using network cascading of the cell-cascaded transmission lines. In this section, a phase compensation method for the cell network cascading is proposed here. The method is good for efficient computation of the phase delay characteristics with acceptable accuracy as compared to the full-wave EM simulation.

As shown in Figure 4.17, these cascading errors can be compensated by the proposed procedure. For low loss condition, the transmission phase delay characteristics are of major concern for both of the transmission lines and the filter designs. Only the phase delay errors are considered in the following method. Assume: the adjacent cells have the same mutual coupling effect, which is treated as the first-order errors Δ . The cells with one interval have the same coupling effect, which is treated as the second-order errors Δ^2 and cells with more than two cells interval have small coupling effects with negligible errors $N(\Delta^2)$.

Figure 4.17 Cell cascaded transmission line with p cells and the errors

The total phase delay of the CBCPW lines can be approximately calculated by adding up the sum of the first-order errors, the second-order errors of the cascaded p T-cells and the phase delay of the directly cascaded p T-cells as

$$\text{Phase}(p * \text{cell}) \approx \text{Cascadephase}(p * \text{cell}) + (p-1) * \Delta + (p-2) * \Delta^2 + N(\Delta^2) \quad (4.14)$$

Δ can be gotten by setting $p=2$ in equation (4.14) as

$$\Delta = \text{Phase}(2 * \text{cell}) - \text{Cascadephase}(2 * \text{cell}) \quad (4.15)$$

Equation (4.15) means that in order to get the Δ , full-wave EM results of the single-cell and two cascaded cells are needed. And Δ^2 can be gotten by setting $p=3$ in equation (4.14) as

$$\Delta^2 = \text{Phase}(3 * \text{cell}) - \text{Cascadephase}(3 * \text{cell}) - 2 * \Delta \quad (4.16)$$

To get Δ^2 , full-wave EM results of three cascaded cells and Δ are needed. By using the above procedure, the absolute values of phase errors Δ and Δ^2 for the lines in Figure 4.10 can be calculated as depicted in Figure 4.18. The absolute value of the first-order errors is much larger than that of the second-order errors. This is mainly because that the adjacent cells have larger mutual coupling effect than the interval cells due to the distances among the cells. Both kinds of errors increase with the frequency. The extracted first-order and second-order errors demonstrate that the traditional cascading network method leads to large errors at higher operating frequency.

Figure 4.18 The first-order and the second-order errors of the cascaded cells

$$R_o=0.26\text{mm}, R_i=0.21\text{mm}, l_i=0.46\text{mm}, W_o=0.1\text{mm}, S_o=0.05\text{mm}$$

Three types of cell lines with nine cells in Figure 4.10 are calculated by the

traditional method of cell cascading, the cell cascading with compensation and the full-wave EM. In Figure 4.19, the phase delay errors due to the traditional cell cascading are compared with that between cell cascading with compensation, using the full-wave EM method as reference. For unloaded line, the phase delay, gotten by cascading with compensation, is almost the same as that of the full-wave EM results. While for the loaded transmission lines, the phase delay errors increase in higher operating frequency. The reason is that the cells with loading have nonlinear phase relationship with the number of the cells. The compensation procedure can mainly reduce the linear phase delay errors of the direct cascading.

Figure 4.19 Comparisons of the phase delay error between direct cascade and full-wave EM results and between cascade with compensation and full-wave EM results

The phase delay errors, between the cascading cells with the phase compensation and the measured phase delay (i.e. the total phase delay of nine cells calculated from errors compensation method minus the measured phase delay of 9-cell lines), are

given in Figure 4.20. The small phase delay errors can be seen in low operating frequency and unloaded transmission line. It can be observed that the total phase delay calculated by the proposed method is more accurate than that calculated by the traditional cascading cell method. Moreover, less computational effort for the proposed method is needed as compared to that for the full-wave calculation of the whole structure. The dip points in Figure 4.19 and Figure 4.20 are mainly due to the phase delay jumps from -3.14 to 3.14 radians (only finite phase delay points with respect to the frequencies are calculated).

Figure 4.20 Phase delay errors between the calculations and measurements

4.3 Loaded and Unloaded CBCPW Filters

4.3.1 Filter design

The configurations of the loaded and unloaded end-coupled quarter-wavelength BPFs are illustrated in Figure 4.21. All these filters are formed by the T-cells and C-cells (nine cells in total) as denoted in Figure 4.21. As an example, the

cell-cascaded topology of the filter in Figure 4.21 b) is demonstrated in Figure 4.22.

At the operating frequency of the filter, the following phase relationship should be satisfied with

$$Phase(p * Cell) = \frac{\pi}{2} + \frac{1}{2} [pha_{j-1,j} + pha_{j,j+1}] \quad (4.17)$$

Figure 4.21 Configurations of second-order CBCPW filters

Figure 4.22 Cell topology of second-order CBCPW filter in Figure 4.21 b)

The $pha_{j,j-1}$ and $pha_{j,j+1}$, which are corresponding to the coupling cell between the resonators $j-1^{th}$ and j^{th} and the coupling cell between the resonators j^{th} and $j+1^{th}$ respectively, can be calculated from equations (4.10) and (4.12). The left hand side of equation (4.17) can be calculated by using equation (4.14). The I/O port coupling

(J -inverter) and the inter-stage coupling (K -inverter) for the filter design can be calculated by following the procedure of a quarter-wavelength filter discussed in [18]. The coupling gap and coupling shunt dimension corresponding to the required K and J values of the filter can be determined by the coupling coefficient extracting procedure given in equations (4.9) and (4.11) in Section 4.1. The design given in Figure 4.21 a) has the same coupling gap $S_I=20\mu\text{m}$, shunt width $W_s=0.1\text{ mm}$ and total length $L=4.38\text{ mm}$ as that in Figure 4.21 b). The design in Figure 4.21 c) has the dimension of $S_I=20\text{ }\mu\text{m}$ and $W_s=0.24\text{ mm}$. The ground vias with a diameter of 0.25 mm are used on each lateral ground with the interval distance of 1.3mm (refer to Figure 4.29). The filter in Figure 4.21 a) is limited by the gap width S_I and the minimum trace W_s of the fabrication process. However, the loaded filters in Figure 4.21 b) and c) can be realized due to the increased coupling coefficient for the same dimensions of the coupling gap and shunt as discussed in section 4.1. The unloaded, IT and CT loaded filters are compared in Figure 4.23. The measured result of the CT loaded filter shown in Figure 4.21 c) has the center frequency of 12.6GHz , the insertion loss of 4.5 dB and the relative bandwidth of 5% . The measured result of the IT loaded filter has the operating frequency of 13.4 GHz and the insertion loss of 3.2 dB . Both types of the loaded filters have good stopband rejection. From Figure 4.23, the loaded filters have lower operating frequencies than the unloaded filter. It means that for the same operating frequency, the loaded filters have smaller size than the unloaded filter. The spectrum of the unloaded and the IT loaded filter from 50MHz to 30.05GHz are compared and shown in Figure 4.24. It can be concluded that the

operating frequency of the loaded filter is mainly determined by the characteristic of the cells and the loaded filters have large coupling coefficients and compact size.

Figure 4.23 Measured results of CBCPW filters with and without loading

Figure 4.24 Measured results of IT loaded and the unloaded BPFs

4.3.2 Discussions on power leakage of the CBCPW filters

In Section 4.2, the power leakage of the CBCPW transmission lines with and without loading can be suppressed to a reasonable level by using either “via ground” or “finite ground”. In this part, we will demonstrate that the CBCPW lines with “finite

ground” can be used to form the leakage-mode-free lines. However, it can not be used to suppress the power leakage in CBCPW structures operating in resonant state like the end-coupled BPFs.

Figure 4.25 Measured results of the CT loaded filters with and without via

Figure 4.26 Measured results of the IT loaded filters with and without via

In Figure 4.25, the measured transmission characteristics of the CT loaded filters with via (with “via ground”) and without via (with “finite ground”) are compared. In

Figure 4.26, the transmission characteristics of the IT loaded filters with and without via are compared. Though the dimensions of the counterparts (with and without via) are the same except the ground vias, the frequency responses of the filters are dramatically different. It can be concluded that the “finite ground” approach, which is suitable for CBCPW transmission lines (refer to Figure 4.16), is no longer acceptable for the end-coupled CBCPW BPFs due to the leakages in the passband and the stopband as shown in Figure 4.25 and Figure 4.26. To further investigate the mechanism, the calculated loss factors $(1 - |S_{11}|^2 - |S_{21}|^2)$ of those filters with and without vias from the measured S-parameters are compared in Figure 4.27 and Figure 4.28. It can be observed that for the filters without ground vias, more loss exists in the lower rejection band close to the passband. The higher loss is due to the power leakages as discussed in Section 4.2. From the results in Figure 4.25 to Figure 4.28, it can be seen that the filters with “finite ground” have much worse stopband characteristics and skirt selectivity in the rejection band compared to the filters with the “via ground”. The main reason is that when the lateral ground is not connected to the backside ground, the input signal can propagate to the output port through the lateral ground with finite width and thus decrease the selectivity of the resonator. The CPM-modes can support these power leakages. However, when the lateral ground is connected to the backside ground, the CPM-modes are suppressed and the direct coupling path through the lateral ground is terminated to the ground. Thus the stopband rejection as well as the skirt selectivity of the filter is improved. It can be seen that the approach of ground vias with proper interval is much better than the

finite ground approach for the leakage-mode-free CBCPW end-coupled filter.

Figure 4.27 Measured loss factors of the CT loaded filters with and without via

Figure 4.28 Measured loss factors of the IT loaded filters with and without via

4.4.3 Size reduction factor of the CBCPW filters

Since the three types of quarter-wavelength CBCPW filters in Section 4.3.2 have the same total length and the number of the filter orders. The total length of the filter is equal to the length of the unit resonator times the number of the resonators (the number of the filter order (denoted as q)). Thus, it is useful to define a size reduction

factor according to the operating frequencies as

$$M \approx \frac{q * \lambda_{g1}/4 - q * \lambda_{g2}/4}{q * \lambda_{g1}/4} = \frac{\lambda_{g1} - \lambda_{g2}}{\lambda_{g1}} = 1 - \frac{f_{01} * \sqrt{\epsilon_{e1}}}{f_{02} * \sqrt{\epsilon_{e2}}} \quad (4.18)$$

where λ_{g1} and λ_{g2} are the wavelengths for the unloaded line corresponding to f_{01} , which is the center operating frequency of the loaded filter, and f_{02} , which denotes the center operating frequency of the unloaded filter, respectively. ϵ_{e1} and ϵ_{e2} are the effective relative permittivity of the transmission lines without loading at two operating frequencies (i.e. f_{01} and f_{02}). Assume the effective relative permittivity for the compared filters at two frequencies is the same, thus

$$M = 1 - \frac{f_{01}}{f_{02}} \quad (4.19)$$

The calculated size reduction factors for the loaded filters shown in Figures 4.21 b) and c) are approximately 23% and 27% respectively. The enlarged photographs of the lines and filters with the via ground are illustrated in Figure 4.29, and the photographs of the lines and filters without via ground are not demonstrated here and the structures are similar except the ground via holes.

Figure 4.29 Enlarged photographs of fabricated lines and filters

Each one with Length: 4.38mm Width: 1.5mm

4.4 Miniaturized Filters Using Zigzag Lateral Grounds

4.4.1 Configuration

In this part, a CBCPW filter architecture based on a zigzag lateral ground is introduced. The filter characteristics, sensitive to the number of via holes, are studied by using HFSS software. The designed second-order BPF operating at 10.6 GHz with 5.4% relative bandwidth has good filtering characteristics in both the passband and the stopband. The zigzag lateral grounds and enhanced end-coupling contribute to filter's size reduction by 24.3% and the filter size is only 5.3 mm by 1.5 mm.

Figure 4.30 Perspective view of zigzag lateral ground CBCPW filter

The CBCPW filter configuration in a perspective view is shown in Figure 4.30. A ground metal layer is under the bottom of the substrate (with the substrate thickness of H_s) and the black CPW metal patterns are built up on the substrate. The structure above the top metal is an air cavity. The top lateral grounds of the CBCPW are connected to the bottom metal ground through the metal via holes. For the BPFs, the power leakages should be suppressed in order to increase the stopband rejection and

to reduce the possible interference to the adjacent circuits as analyzed in section 4.3. Therefore, via holes with a proper interval spacing are needed in the filter design. The pattern of the top metal with footprints of four-pair via holes is illustrated in Figure 4.31 a). The lateral grounds with width of W_g are defected with triangle shapes periodically. The filter is formed by eleven cells (eight transmission cells (T-cells) and three coupling cells (C-cells) discussed in Section 4.3). The length of L_c for each cell is illustrated in the filter configuration in Figure 4.31.

a) with zigzag lateral ground

b) with uniform lateral ground

Figure 4.31 Top metal of the CBCPW filters

As a contrast, the similar BPF with the same total length but without zigzag lateral ground and coupling loading is shown in Figure 4.31 b). The design procedures given in Section 4.3 are used in the initial design. For the CBCPW filter, the filter characteristics such as inter-stage coupling, port coupling and radiation loss are affected by the lateral grounds, the number of via holes and the dimension of the

substrate. The full-wave EM software HFSS is used in the following design and analysis. The simulation results of the filters in Figures 4.31 a) and b) are compared in Figure 4.32. The operating frequencies of the filter in Figures 4.31 a) and b) are 10.6 GHz and 14 GHz respectively. The calculated size reduction factor for the loaded filter shown in Figure 4.31 a) is approximately 24.3%.

Figure 4.32 Simulation results of the filters in curves a) and b) of Figure 4.31

Zigzag: Figure 4.31 a) Uniform: Figure 4.31 b) Dimensions: $D=0.25$ mm, $G=0.1$ mm, $W_s=0.12$ mm, $L_c=0.46$ mm, $W_o=0.1$ mm, $S_o=0.05$ mm, $W_g=0.64$ mm

The experimental and simulated results are compared in Figure 4.33 and Figure 4.34. Both results have matched well with each other. The measured insertion loss of 2.9 dB is 0.9 dB larger than that of simulated 2.0 dB. Deep rejections are achieved in higher and lower stopband. Rejection of 20 dB is achieved at frequency points of 1 GHz offsetting from center operating frequency of 10.6 GHz. The measured return loss is better than 20 dB. It is noted that the second-order filter designed in [25] has almost the same operating frequency as the designed filter. The specifications are

compared in Table 4.1. Better performance and compact size are achieved by using the proposed filter.

Figure 4.33 Simulated and measured results of the filter in Figure 4.31 b)

TABLE 4.1 PERFORMANCE COMPARISONS OF THE FILTERS

	Type	Operating frequency	FBW	substrate	IL/RL (dB)	Size (mm ²)
[25]	MCS line	10.5 (GHz)	4%	Alumina $\epsilon_r = 9.9$	3.4/-11	>15 by 4
This filter	CBCPW line	10.6 (GHz)	5.4%	Alumina $\epsilon_r = 9.8$	2.9/-22	5.3 by 1.5
Note: FBW (relative bandwidth), IL (insertion loss), RL (return loss)

Figure 4.34 Results of the filter with 4-pair metal via holes in narrow band range

4.4.2 Discussion on lateral grounding effects

The boundary conditions such as via holes and the lateral ground can greatly affect the filter performance. The filter in Figure 4.31 a) with shielding box is shown in Figure 4.30. The box dimensions are $W=4.34$ mm, $L=5.3$ mm and $We=1.48$ mm. The height of the box is chosen to be 2.4 mm, which is about six times of the substrate height H_s . The top cover effects can be neglected. Three cases of the shielding are: case X: four pair of via holes in lateral ground as the top pattern in Figure 4.31 a); case Y: the via holes are not used and the outside of lateral grounds are connected to backside (i.e. $W=We=1.48$ mm) ground using shielding metal walls; case Z: both the via holes or shielding walls are not used. The perfect conductor is assumed for both metal via and the shielding walls. The EM simulation results of the three cases are compared in Figure 4.35. The transmission characteristics of case X and Y are almost the same except at the operating frequency of case Y which is slightly higher than that of case X. This demonstrates that proper metal via holes can be used to replace the

metal ground wall. But case Z is completely different from cases X and Y, due to the overmoded excitation of the finite lateral ground CBCPW. The increase of the number of via holes can reduce high-order modes generated in the higher operating frequency range. The photographs of two filters with 4-pair via holes and 5-pair via holes are shown in Figure 4.36.

Figure 4.35 Comparison of filters with different shielding types

X: 4-pair metal via holes Y: shielding walls in lateral ground Z: without shielding walls and via holes

Figure 4.36 Enlarged photographs of filters

Each one with Length: 5.3 mm (208.7mil) Width: 1.5 mm (59mil)

4.5 Conclusion

The CBCPW transmission lines and filters with/without periodic loading are implemented using thin-film technology. The compact size, narrow band, coupling

enhancement and steep roll-off BPFs are achieved using the CBCPW for the first time. The procedures of the model-based EM extracting and cascading of unit cell are demonstrated and used for the filter design. A compensation method for directly cells cascading is introduced to improve the phase delay accuracy for directly cascading cells. It is demonstrated that the “finite ground” approach for free of power leakage is not applicable to the resonant circuits like end-coupling filters. It is also demonstrated that the “via ground” approach can overcome the shortcoming of “finite ground” CBCPW in the resonant circuits. The approach can reduce the insertion loss and improve the skirt selectivity and the stopband rejection. These investigations should be helpful for design and packaging of the CBCPW circuits. These types of filters and lines, with advantages of compactness and uniplanar configuration, are attractive to be used in the MIC/MMIC circuits.

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CHAPTER 5

FILTER DESIGN USING SILICON-BASED MICROMACHINING TECHNOLOGY

Thin-film process on ceramic substrate in Chapter 4, which can fabricate narrow metal traces and gaps among metal traces down to 20 μ m, demonstrates a much better process capability for filter miniaturization as compare to PCB technology in Chapter3, which currently can only handle the narrowest trace and gap to 100 μ m. Recently, the accuracy of silicon-based micromachining technology can scale the devices down to several micrometers and even below hundreds of nanometer. Silicon has also many advantages as a substrate which includes low cost, good thermal [1] and mechanical [2] properties as well as mature manufacture technology. The ability to chemically “micromachine” with precise grooves and holes in silicon has been utilized to produce microwave and millimeter wave components [3]-[11].

In this chapter, some work on resonators and filters using silicon micromachining technology is presented. Several devices are designed and fabricated by using SiDeox (silicon deep etching and oxidation) and through-hole plating technology [12]. The main focus is on designing and analyzing some special resonators and filters. In Section 5.1, the design and characterizing of a CBCPW feed ring resonator, which is suitable for device mounting, are introduced. In Section 5.2, by using the transition from the CBCPW to coupled MS lines, a high pass filter with compact size is realized. A micromachined ground ring guarded path resonator/filter

is presented in Section 5.3. A miniaturized ground ring guided patch filter with ground shunts, which is compatible with CBCPW feeding structures in I/O ports, is proposed and investigated in Section 5.4. The proposed ground ring guarded patch filters with ground shunts is 30% the size compared to the traditional patch filter.

5.1 CBCPW Feed CBCPS Ring Resonator

5.1.1 General ring circuit

The ring resonators have been widely used to measure the effective dielectric constant, dispersion, phase velocity and discontinuity parameters, to determine the optimum thickness of the substrate thickness [10], [11],[13] and to build up filters, oscillators, mixers and antennas [14]-[16]. The advantages of the ring resonators include the freedom from the open-end effect, compact size and low radiation loss. At the same time, the curvature effect of the ring resonators can be made negligible if the diameter of the ring resonators is large enough with respect to the line width [10]. Though MS line is the most popular planar transmission line, other transmission lines such as CPW, coplanar strips (CPS), waveguide, and slotline are alternatives to MS line in microwave and millimeter wave circuits. The CPS, a complement to the CPW, supports all advantages of the CPW and is useful in manufacturing transmission lines with higher characteristic impedances and smaller size than the CPW [9],[17],[18]. Furthermore, due to the balanced configuration of CPS, CPS is also useful for mounting the active devices such as diodes to form the tunable circuits.

Figure 5.1 The top view of the CBCPW feed CBCPS ring resonator

In this part, a CBCPW feed conductor backed CPS (CBCPS) ring resonator as shown in Figure 5.1 is proposed and implemented. The ring is fabricated on the silicon wafer and measured by probe station. The characteristic of the proposed ring resonator are also compared with that of the MS ring resonator.

5.1.2 Ring resonator description

The conductor backing leaves the CPW circuit susceptible to the power leakage into surface waves or into the region between the conductor plates [17]-[21] as shown in Chapter 4. In order to decrease the loss of silicon substrate and reduce the leakage power of CBCPWs, the silicon wafer is thermally oxidized on both sides to form three dielectric layers $\text{SiO}_2\text{-Si-SiO}_2$ as shown in Figure 5.2. Furthermore, to ensure that the transmission lines support only a quasi-static TEM mode, two ground plates on the topside of CBCPW are connected to the bottom ground plate by metal vias. The metal vias are used to short-circuit the electric fields of the parasitic parallel-plate modes and suppress their propagation. The experimental results show that power leakages can be controlled by properly arranging of the vias as shown in Chapter 4.

Figure 5.2 Cross section of three-layer dielectric transmission lines

The CPCPS line as described above is used to form a closed loop on the substrate as shown in Figure 5.1. Without consideration of the curvature effect, the resonant frequencies can be determined by assuming that the structure will support only waves that have an integral multiple of the wavelength equal to the mean circumference of the ring slot. At the resonant frequency, it can be expressed as [10][22]

$$2\pi R_c = n\lambda_g \quad n = 1, 2, 3, \dots \quad (5.1)$$

$$f_r = \frac{nc}{2\pi R_c \sqrt{\epsilon_{eff}}} \quad (5.2)$$

where λ_g is the guided wavelength, R_c is the mean radius of the slot, n is the mode number, f_r is the resonant frequency, c is the speed of light in free space, and ϵ_{eff} is the static effective relative permittivity.

The CBCPS ring is excited by the weak coupling for the convenience of extracting of the quality factor [10]. As shown in Figure 5.1 and Figure 5.2, the via-CBCPW transmission line with 50Ω characteristic impedance is used in the input and output port. The transition between different physical dimensions of via-CBCBW transmission lines, which have the same characteristic impedance (50Ω), is to meet the requirement of probe-tip test and gap coupling. The mean radius and width of the slot

between the inner and outer ring conductor are $R_c=1.88$ mm, $S=0.12$ mm respectively. The width of inner ring conductor is $W_i=0.23$ mm and width of outer ring conductor is $W_o=0.4$ mm. To investigate the difference between the CBCPS ring resonator and the MS ring resonator, the MS ring is also built. Under the same via-CBCPW fed lines, coupling gaps $G=0.06$ mm, the spectrum characteristics of MS ring (mean radius of the MS ring $R_m=1.73$ mm, conductor width of the MS ring $W=0.23$ mm) and the CBCPS ring are compared in Figure 5.3 by using the EMSight from AWR.

Figure 5.3 Comparison of the MS and CBCPS rings in narrow frequency range

The MS ring is resonant at the frequency of $f_m=9.83$ GHz with bandwidth of $f_{3dB}=80$ MHz, insertion loss of -9.8 dB at f_m . While the CBCPS ring is resonant at frequency of $f_r=10.2$ GHz with $f_{3dB}=75$ MHz, insertion loss 10.2 dB at f_r . The loaded Q-factor can be calculated by

$$Q_L = f_r / f_{3dB} \quad (5.3)$$

where f_r is resonant frequency and f_{3dB} is 3-dB bandwidth of the resonator. The Q_L values of the MS ring and CBCPS ring are 122.9 and 136 respectively. The practical

value of the Q-factor is smaller than the calculated value, because the metal conductor loss is not considered in the simulation.

Figure 5.4 Comparison of the MS ring and CBCPS ring in wide frequency range

The spectrum characteristics of the MS ring and the CBCPS ring are compared in Figure 5.4. The first harmonics of $n=2$ mode of MS ring resonator and CBCPW ring resonator are 18.9 GHz and 19.8 GHz respectively. These two rings have similar spectrum response. There exist the weak parasite resonant phenomena in the spectrum of the CBCPS ring resonator. The parasite resonances in 8.15 GHz and 17.3 GHz marked as $Pr1$ and $Pr2$ respectively are due to the leakage modes [17]. Because the parallel plate (PP) mode in the CBCPW is suppressed (refer to following experimental demonstration of the transmission line), the parasite mode should be generated by the CBCPS. Part of the leakage power of PP modes in the CBCPS transmission line re-enters into the CBCPS transmission line due to the confined structures of the CBCPS ring. This leads to the dip points or the parasitic resonance in the spectrum response. Fortunately, the PP mode is weak enough. The current distribution on the top conductor of CBCPS ring is illustrated in Figure 5.5. The current directions between the inner

conductor and outer conductor strips of the CBCPS ring are reversed as denoted by the black arrows.

Figure 5.5 Current directions of the CBCPS ring resonator

Figure 5.6 Measured spectrum response of Via-CBCPW line

Figure 5.7 Measured spectrum response of CBCPS ring resonator

Figure 5.6 illustrates the measured results of a section of via-CBCPW line with longitude length of 5.85 mm and the same transverse dimension of the line as the feed line of the ring resonators. No PP mode is excited in the investigated frequency range. The measured spectrum response of the ring resonator in Figure 5.1 is demonstrated in Figure 5.7. The minimum insertion loss is -23.2 dB. Under weak gap coupling condition, the feeding structures including the I/O ports have weak effect on the unload Q factor and the resonant frequency. Weak coupling in I/O ports is good for accurately extracting of these parameters. The measured resonant frequency f_c and f_{3dB} of the CBCPS ring are 10.62 GHz and 220MHz respectively. The measured $Q_{L,measure}$ is about 48.3 by using equation (5.3) and the measured unloaded Q of the resonator can be obtained by

$$Q_{u,measure} = \frac{Q_{L,measure}}{1 - 10^{-L_{measure}/20}} \quad (5.4)$$

where $Q_{L,measure}$ is the measured loaded quality factor, $L_{measure}$ is the measured insertion loss in decibels of the resonator at resonance.

The measured unloading Q-factor of fundamental mode ($n=1$ resonant mode) of the resonator is 53.5. The unloaded Q-factor of the $n=2$ resonant mode is 45. Because of the weak coupling in I/O ports, the value of the loaded Q-factor for the fundamental mode is quite close to that of the unloaded Q-factor. Because the thin SO_2 layers with low permittivity (traditionally is about 4) decrease ϵ_{eff} , the resonant frequency increases by 4.1%. The higher insertion loss of the experiment results compared with that of the simulation results is mainly due to the thickness of the top conductor (measured sheet resistivity is $0.078\Omega/\square$). The thin copper with thickness of $0.6\text{ }\mu\text{m}$ formed by sputtering machine reduces the coupling coefficient between fed-CBCPW line and the CBCPS ring and also leads to more metal loss in the lower operating frequency range (the skin depth of copper is $0.66\text{ }\mu\text{m}$ at 10 GHz). The degradation of the resistivity of the silicon during fabrication as interpreted in [18] also increases the loss.

5.2 A Planar Miniaturized High-Pass Filter

5.2.1 Circuit description

In this part, a miniaturized silicon based high-pass filter, using CBCPW with ground via (via-CBCPW) cascaded with three coupling MS lines (called HCTL), is realized. The design of the filter makes use of the transition from the CBCPW to the coupled MS lines, by which the necessary coupling for the filter design is also achieved. Therefore, the small size can be achieved by omitting the extra length for the transition from CBCPW to the MS. The via-holes for the CBCPW ground can also

be modeled as the lumped inductors, which have the loading effects and contribute to size reduction of the filter (refer to Chapter 2 and Chapter 4). Due to the mode transition and multiple coupled MS lines, the additional ZPs and pole points are generated. The ZP in the stopband increases the skirt selectivity in the stopband and the additional poles extend the passband bandwidth in the upper passband. The top view of the proposed high pass filter with HTCL structure is shown in Figure 5.8 a). In Figure 5.8 b), H denotes the thickness of the silicon ($H=0.5\text{mm}$, resistivity of $4000\Omega/\square$ before fabrication) and SiO_2 layer is thermally oxidized on the Si to reduce the loss of the substrate. The circuit is composed by three parts: the coplanar waveguide L_{cop} , the hybrid coupling transmission lines L_{hy} and the MS of L_{ms} . Due to the symmetry in the input and output ports of the circuit, only half of the circuit is denoted.

Figure 5.8 Proposed HCTL high pass filter

a) top view b) vertical partial cutting

5.2.2 Filter results

As shown in Figure 5.8, the Via-CBCPW is used in the I/O port. In this Via-CBCPW circuit part, there is a 50Ω to 50Ω transition with the length of L_{tr} which realizes the transition from the internal 50Ω transmission line with the dimension of $W_1=0.2$ mm and $S_1=0.12$ mm to 50Ω transmission line with $W_2=0.1$ mm and $S_2=0.05$ mm for the RF-probe measurement (here W_1 and W_2 are center conductor width and S_1 and S_2 are the distances from the center conductor to the side conductor). By properly choosing the lengths of L_{ms} , four poles (two are at 12 GHz and the other two are at about 22 GHz) in the passband and one zero point (at about 11 GHz) in the low stopband are generated as demonstrated in Figure 5.9. The physical dimensions of the high pass filter are: $H_t=2.53$ mm, $L_t=3.7$ mm, $L_{hy}=1.36$ mm, $L_{tr}=0.36$ mm, $L_v=0.55$ mm, the radius of the via $R_{via}=0.05$ mm, the distance of $D_{via}=0.2$ mm between the center of the two adjacent vias on the same side conductor of CBCPW. The proposed filter has smaller size compared to the reported structures for the planar microstrip filter under the same kind substrate and operating frequency. The measured results are shown in Figure 5.10. The simulation results and the test results are agreed well in the trend and the number of the zero and poles.

Figure 5.9 Simulation result of the HCTL high pass filter

Figure 5.10 Measured results of the HCTL high pass filter

5.3 A Ground Guarded Slit Patch Resonator

5.3.1 General patch circuit

In the past, as a fundamental building block for filters, the resonator has been extensively studied [22]-[27]. MS patch resonators [22], which have different shapes such as triangular, circular, etc., are of interest for the design of MS filters to increase the power handling capability [22],[23]. MS patch resonators also have lower conductor loss as compared with MS transmission line resonators. However, the patch

resonators and filters tend to have stronger radiation loss to the space as well as to the substrate. It is also possible for these radiations to generate interference to nearby circuits. Therefore, a metal housing is needed to minimize the radiation [22] and interference.

As introduced before, the CBCPW [28],[29] has extra advantages such as higher heat sinking ability, stronger mechanical strength, and lower characteristic impedance. It is attractive to be used in MMIC and packaging. In this part, a novel resonator/filter without metal housing is introduced to achieve lower radiation loss, lower operating frequency and more compatibility with the CBCPW feed structures as compared to traditional patch resonator.

5.3.2 Ground guarded patch resonator with slits

The ground guarded slit patch resonator/filter is shown in Figure 5.11. The traditional MS rectangle patch resonator with width of W and length of L is reshaped to the slit patch resonator as shown in Figure 5.11. The lateral grounds of the via-CBCPW (see sections of L_f and L_{tr}) in I/O ports are extended to surround the slit patch and to keep the slot width of S_c between the ground-guard ring and the patch. The lateral grounds, which surround the patch, form the ground-guarded ring. In the via-CBCPW parts of I/O feeders, there is a 50Ω to 50Ω transition with the length of L_{tr} , which realizes the transition from internal 50Ω line with length of L_f to external 50Ω line with narrow metal trace and slot for the RF-probe tip measurement. Ground vias, connecting the lateral ground to the backside ground, are used to suppress the

possible leakages through the substrate. The ground-guarded ring can shift up the operating frequency of the patch a little due to the fringing field at the edge of the patch resonator.

a) Perspective view of the configuration in HFSS

b) The top metal pattern

Figure 5.11 Configuration of the ground guarded slit patch resonator/filter

To demonstrate the effect of the ground-guarded ring and slits on the patch resonator, three configurations using the same via-CBCPW I/O parts ($W_a=0.2$ mm, $S_a=0.12$ mm, $L_f=1.21$ mm, $L_{lr}=0.36$ mm, $D=0.2$ mm, $W_d=0.04$ mm, $S_d=0.08$ mm) are compared using the full-wave EM simulation: a) the traditional rectangle patch ($W=2$ mm, $L=1.96$ mm); b) the slit patch ($W_p=0.76$ mm, $L_p=0.84$ mm); c) the slit patch with ground guarded ring ($W_b=0.12$ mm, $S_b=0.08$ mm, $S_c=0.12$ mm, $W_e=0.2$ mm). The loss

factors $(1 - |S_{11}|^2 - |S_{21}|^2)$ and unloaded quality factor Q_u of the three structures, calculated with HFSS, are compared in Figure 5.12.

Figure 5.12 Comparison of the response of three patch resonator/filter

The resonator is resonant around the frequency where the loss factor is the maximum. Since the substrate, the metal and the boundary of the three configurations are the same in the simulation, it is clear that the ground-guarded slit path has the lowest operating frequency (13.8 GHz), the smallest maximum loss factor (0.28) and the highest unloaded Q -factor (44.7) among the three configurations. Compared with traditional patch, which operates at 19.4 GHz with loss factor (0.41), the ground guarded slit patch can contribute to size reduction, radiation loss reduction and Q -factor improvement.

Figure 5.13 Comparison of the simulation and measurement

Figure 5.13 shows the comparison of the simulated results from HFSS and the measured results. Satisfactory agreement has been achieved in both the passband and the stopband. The resonator/ filter is operating in 13.8GHz with -3 dB bandwidth of 1.62 GHz, insertion loss of 1.36 dB and return loss of about -20 dB in the passband. The measured insertion loss is a little smaller than the simulated 1.74 dB. The measured unloaded Q-factor of 58.8 is larger than that of the simulated 44.7. This difference is due to the thin oxide layers. The thin oxide layers (See Figure 5.2), which can reduce the substrate loss up to 0.5 dB [6] and improve the Q-factor, are not considered in the simulation in order to reduce computation. The size is only about $5.6 \text{ mm} \times 2.65 \text{ mm}$ as the circuit photograph shown in Figure 5.14.

Figure 5.14 Realization of the filter

5.4 Ground-Guarded Patch Resonator and Filter

It is well known that the performance and the size of filters are mainly determined by the characteristics of resonators when the topology of filters is determined. Traditional patch resonators have a large size, which is demonstrated as a disadvantage and limits its utilization in the high density integrated circuits. In this part, a miniaturized ground-guarded patch resonator and filter, which is compatible with direct CBCPW feeding structures in I/O ports, is proposed and investigated. Under the same circuit dimensions, the operating frequency for the resonator/filter with the ground shunt is only $1/3$ of that for a traditional patch resonator/filter without the ground shunt. The proposed filter as shown in Figure 5.15 is realized on silicon wafer by using micromachining technology.

Figure 5.15 Perspective view of ring guarded shunt patch filter

■ Metal

5.4.1 Guarded patch resonators

As shown in Figure 5.16 a), the patch resonator, with width of W_p and length of L_p , is surrounded by the metal ground-guard ring which is directly connected with lateral ground of the input CPCPW transmission line (center strip width of W , slot width of S_{in}). A section of vertical metal trace with width of W_l , and the center traces of CBCPW in I/O feed network form a “T” shape feed-in structure. The ground-guard ring can shift the operating frequency of the patch resonator up a little with increased via number, by changing the fringing field at the edge of the patch resonator. The resonant frequency [30], [31] for mn^{th} mode could be obtained by

$$f_{mn} = \frac{c}{2\pi\sqrt{\varepsilon_{dyn}}} \sqrt{\left(\frac{m\pi}{L_{eff}}\right)^2 + \left(\frac{n\pi}{W_{eff}}\right)^2} \quad (5.5)$$

where ε_{dyn} is the dynamic permittivity of the patch defined in [30], c is the speed of light in free space, L_{eff} and W_{eff} are effective width and length respectively [31].

Figure 5.16 Ground ring guarded patch resonators

a) rectangular b) rectangular with shunt c) “U” shape d) “U” shape with shunt

■ Substrate ▨ Metal

The ground ring guarded patch resonator with ground shunt is shown in Figure 5.16 b). The only difference between the structures of Figure 5.16 a) and Figure 5.16 b) is that a ground shunt with width of W_s is used to connect the patch to the ground-guard ring in the structure of Figure 5.16 b). In Figures 5.16 c) and d), the patches are “U” shape with and without shunt respectively and the center strip of the CBCPW feed line is extended by the length of L .

As a comparison, the frequency response characteristics of the structures shown in Figure 5.16 are presented in Figure 5.17. The resonant frequencies of the resonator in Figure 5.16 a) are 20.7 GHz ($m=1, n=0$) and 21.3 GHz ($m=0, n=1$) using equation (5.5), while the resonant frequencies of the resonator in Figure 5.16 b) are 5.9 GHz and 18.9 GHz. Obviously, for the same dimensions, the lowest resonant frequency of the ground-guarded resonator in Figure 5.16 b) is only about 30% that of the resonator without ground shunt. For the two “U” shape resonators in Figure 5.16 d) and c), the lowest operating frequencies are 5.5 GHz and 18.7 GHz respectively. Thus for the same operating frequency, the ground shunt can contribute to the size reduction of

more than three times as compared to the patch resonator without ground shunt.

Figure 5.17 The frequency response of the resonators in Figure 5.16

$W_l=50$ μm , $W_p=2$ mm, $L_p=2.28$ mm, $W_s=0.2$ mm, $S=0.15$ mm, $W_l=0.1$ mm, $S_l=0.1$ mm, $W_g=0.3$ mm, $L=1.8$ mm, $W=0.2$ mm, $S_{in}=0.12$ mm

5.4.2 Guarded patch filter with ground shunt

Four second-order patch BPFs are constructed by using the four types of resonators as shown in Figure 5.16. Since we use the similar filter architecture, only the filter built by two resonators in Figure 5.16 d) is demonstrated in Figure 5.15. The I/O ports coupling is mainly determined by the line width of W_l and coupling gap of S_l , while the extended feed line with length of L and width of W can further increase the coupling. The inter-stage coupling is mainly determined by the gap width G between the two patch resonators as shown in Figure 5.15. The procedure in [22] is adopted to design the ground-guarded patch filter with and without ground shunt. In Figure 5.18, the frequency response characteristics of the second-order filters with Figure 5.16 a)

and Figure 5.16 b) type resonators are compared. The resonant frequencies of the filters with and without ground shunt, as denoted by arrows, are 5.5 GHz and 20.5 GHz respectively. The filter with shunt also has wider rejection band width and steeper rolloff.

The filter in Figure 5.15, with 11% relative bandwidth, is designed and the shielding effect of the metal housing with different height of H is compared as shown in Figure 5.19. Without the metal housing, the insertion loss is 3.4 dB, which is much larger than that with the metal housing (the insertion loss is 1.7 dB when the housing height $H=3$ mm). When H is below 3 mm (six times of the substrate height $H_{sub}=0.5$ mm), the decreased height H can shift the operating frequency up. When H is larger than 3 mm, the increased height H almost has no effect on the frequency response.

Figure 5.18 Comparison of Figure 5.16 a) and Figure 5.16 b) type filters

$A=5$ mm, $B=7.9$ mm, $H=5$ mm, $S_f=0.05$ mm, $G=1$ mm, (see Figure 5.17)

Figure 5.19 The shielding effects on ring guarded shunt patch filter

The measured results and the theory results of the filter type in Figure 5.15 are compared in Figure 5.20 and Figure 5.21. As shown in Figure 5.20, the theory and experiment results are in good agreement. The simulated and measured loss factor $(1 - |S_{11}|^2 - |S_{21}|^2)$ of the filter are compared in Figure 5.21. The filter losses are composed of substrate loss, metal loss as well as the radiation loss. Since the wafer is measured without metal housing, the radiation loss is about 1.7 dB in the passband as referred to the metal housing effects demonstrated in Figure 5.19. That means if the filter is properly shielded with metal housing, the measured insertion loss of 4.2 dB can be reduced to about 2.5 dB. The photograph of the filter is shown in Figure 5.22. The size of the filter is only 8.7 mm x 2.9 mm.

Figure 5.20 Comparison of the theory and experiment results of the proposed filter

Figure 5.21 Comparison of the theory and experiment loss factor of the filter

Figure 5.22 Enlarged photograph of the proposed filter in Figure 5.15

5.5 Conclusion

In this chapter, the proposed CBCPS ring resonator and ground ring guarded patch resonators and filters with/without ground shunt are investigated extensively. The novel ground ring guarded patch resonators and filters with ground shunt, which have much compact size and deep stopband rejection, are implemented for the first time. The ground-guarded technique is useful to reduce the radiation loss and improve the stopband rejection of the filters. These MEMS devices are fabricated by the author with the support of MEMS center in Nanyang Technological University.

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CHAPTER 6

RESONATOR AND BALUN DESIGNS USING THE STANDARD CMOS TECHNOLOGY

Both the silicon-based micromachining technology, as introduced in Chapter 5, and commercial CMOS technology can scale the devices down to nanometer range, which is good for miniaturization as well as high accuracy for fabrication. However, silicon-based micromachining technology demonstrates higher cost, less popular and less compatibility as compared to the CMOS technology. As introduced in Chapter 1, CMOS technology is one of the most interesting processes among today's popular processes especially for wireless communications due to its maturity, low cost and compatibility with digital circuits as well as analogy circuits [1]-[4]. In this chapter, some distributed passive circuits with much compact size using standard CMOS technology are presented. Two kinds of passive circuits, dual microstrip meander line (DMML) ring resonator and multilayer stacked balun are proposed and developed. A Ka band DMML ring resonator is presented in Section 6.1. A miniaturized balun is introduced and investigated in Section 6.2. The parasitic effects of the baluns are also investigated. Asymmetrical configuration and tuning capacitors are introduced to further improve the performance of the balun.

6.1 Ka Band CMOS DMML Ring Resonator

The traditional annular ring resonator cannot be economically implemented on

the standard CMOS process due to its huge size especially in a lower operating frequency. In this part, a DMML ring resonator is designed and implemented by using the 0.18 μm standard CMOS process provided by Chartered Semiconductor Manufacturing Ltd (CSM).

Figure 6.1 CMOS dual-meander lines ring resonator

a) top view b) 3D view

The 0.18 μm CMOS process of CSM has six metal layers from the metal one (bottom metal) to the metal six (top metal) on the P-Si-Substrate. The structure is shown in Figure 6.1. The top conductors of the DMML ring, which include meander line 1, meander line 2, the “guard ring” and two “T” type feeders, are built by using the top metal of the CMOS process. The conductors around the DMML ring and feed

structures are connected to the metal one (bottom metal in Figure 6.1 b) of the CMOS process by using the metal vias. Thus the “guard ring” structure, which is used to reduce the crosstalk to neighbor circuits [12]-[13], is formed. As demonstrated in Figure 6.1 b), the metal layers between the top metal and metal one of CMOS process are not employed. The space between the top metal and metal one is the low loss oxide substrate. The high loss P-Si-Substrate is under metal one.

6.1.1 Dual-meander line ring

The meander line as shown in Figure 6.2 is often used to adjust the skews among several signal lines in the synchronous high speed system. It serves as the impedance transformer [11] and forms the inductors in integrated circuits [8] [14]. The primary concern for this study is to use meander lines to reduce the size of the annular ring resonator. To design the DMML ring, the meander line should be designed firstly.

Figure 6.2 Top view of the meander line

As illustrated in Figure 6.2, the width of the section lines W , the gap between two

adjacent section lines S and the length of the section lines L mainly determine the characteristics of the meander line when the substrate and metal parameters are fixed.

So the total physical length of the meander line can be determined by

$$L_{meander} = m \cdot (L + W) + (m - 1) \cdot S \quad (6.1)$$

where m is the number of section line. When L , W and S are fixed and the dispersive characteristic of meander line is ignored, the electrical length of the meander line is linearly related with m . If we assume total oxide with a relative permittivity of 4.0 and thickness of $6.7 \mu\text{m}$, the transmission characteristics of the magnitude and phase are illustrated in Figure 6.3. The best matching point is at the frequency point, where the insertion loss is the smallest and the phase delay is about a half-wavelength or 180 degrees. The best matching frequency point in Figure 6.3 is at 27.6 GHz.

Figure 6.3 The transmission characteristics of meander line

Note: the total oxide thickness of $6.7 \mu\text{m}$ without P -silicon substrate under metal one, $W=5 \mu\text{m}$, $S=5 \mu\text{m}$, $L=195 \mu\text{m}$ and $m=22$, Relative permittivity of oxide=4

Since the meander line is a kind of slow-wave transmission line below certain frequency range [10], if the meander lines are used to replace the transmission line in annular ring, the ring with meander lines should have similar characteristics as the annular ring below certain frequency range. The resonating frequency of the DMML ring resonator is at equivalent integer wavelength as

$$2L_{meander} = n \cdot \lambda_g \quad n = 1, 2, 3, \dots \quad (6.2)$$

$$f_r = \frac{nc}{2L_{meander} \sqrt{\epsilon_{eff}}} \quad (6.3)$$

where λ_g is the guided wavelength of the meander line, $L_{meander}$ is the mean length of a single meander line, n is the mode number, f_r is the resonant frequency, c is the speed of light in free space, and ϵ_{eff} is the static effective relative dielectric constant of the meander line.

Two meander lines shown in Figure 6.2 are cascaded end to end to form the DMML ring resonator as illustrated in Figure 6.1. According to equations (6.2) and (6.3), if the whole resonator operating at the frequency of the fundamental mode $n=1$, the electric length of meander line 1 and 2 should be an equivalent half-wavelength. The I/O feed networks employ the CBCPW structure with physical dimensions: centre conductor width of 50 μm , the gap width (between centre conductor and finite lateral ground) of 30 μm . The coupling between the DMML ring and the I/O port is enhanced by using this edge coupled “T” type feeders (see Figure 6.1). The “T” feeders are formed by the center conductor of CBCPW in I/O feed networks and a section of MS line. In order to reduce computation, the P-silicon substrate under the metal one is not considered during the EM simulation. The EM results are

demonstrated in Figure 6.4. The operating frequency of 27.4 GHz for the DMML ring and the half-wavelength frequency point of 27.6 GHz for single meander line exactly verify the equations of (6.2) and (6.3) when n is equal to 1. The operating frequency shift of 0.2 GHz is possibly due to the connections between the two meander lines and the coupling effect between the meander lines and the “guard ring”.

Figure 6.4 Simulated spectrum response of DMML ring

6.1.2 Realization and measurement

To realize the DMML ring resonator on CSM 0.18 μm standard CMOS process, the top metal with the thickness of 2 μm is used to form the top pattern of the DMML resonator. The metal one above the P-Silicon-Substrate ($\rho \approx 15 \Omega \cdot \text{cm}$) is used as the suspended MS ground. The DMML ring resonator is on the top of the P-Silicon-Substrate.

Figure 6.5 The DMML ring resonator under test

The Agilent 8510C network analyzer, HP8516A S-parameter test set, HP83623A spectrum synthesizer and 50A-GSG-100-VP microprobes from CASCADE MICROTECH are used to fulfill the measurement of the DMML ring resonator. Microwave calibration is taken from the stored values in the analyzer, which are made with the standard short-open-load-through measurements on a GGB CS-5 calibration substrate using the same cables and probes. The device under test (DUT) is illustrated in Figure 6.5. The measured results are shown in Figure 6.6. The measured resonant frequency f_r and 3-dB bandwidth f_{3dB} of DMML ring are 27.85GHz and 5.69GHz respectively, and the minimum insertion loss is 16.15dB at the resonant frequency. The calculated unloaded Q factor of the DMML ring resonator is 5.79 at the center operating frequency 27.85 GHz.

Figure 6.6 Measured spectrum response of DMML ring

By comparing Figure 6.4 and Figure 6.6, it can be seen that the spectrum responses of the DMML ring resonators with and without *P*-silicon substrate under metal one are different. But the resonant frequency of the DMML ring resonator with silicon is almost the same as that of the DMML ring resonator without silicon. The Q factor of the DMML ring with *P*-silicon-substrate is much smaller than its counterpart (Q factor 97.8). The poor skirt selectivity in interested spectrum and large insertion loss in the center frequency are due to the low resistivity of *P*-Silicon-Substrate. Even though the ground of the DMML ring resonator is built on the metal one, the low resistivity silicon beneath ground still increase the loss of the resonator. From the above analysis and experiment, the equations (6.1) and (6.2) can be used to predict the operating frequency of the DMML ring with acceptable accuracy. The effect of the *P*-Silicon-Substrate on the loss and Q -factor of the resonator is dominant, and the effect on the operating frequency is quite small.

6.2 Design of Miniaturized Stacked Baluns on CMOS

To get better performance, differential architectures are commonly used in wireless communication transceivers [1]-[2], [15]. Since in most cases, the received signal from an antenna is a single-ended signal, a balun is widely adopted to provide balanced differential outputs from the single-ended input signal. There are numerous publications on various baluns in the literatures [16]-[24]. However, those reported single-ended drive baluns are not realized on standard CMOS technology. In [23], the single-ended drive baluns are realized on high resistivity silicon ($>4\text{K}\Omega \cdot \text{cm}$). But the balun architectures [23] can not operate in the high loss substrate ($\rho \sim 15\Omega \cdot \text{cm}$) as demonstrated by the authors of [23]. The CMOS transformer and trifilar balun with lower insertion loss and broader bandwidth are achieved from differential drive in [24]. However, the differential drive of trifilar balun still needs the differential drive signal in its' input port. This means the single drive balun is necessary. On the other hand, the trend of Tx/Rx front-ends in wireless communication is to have as less off-chip components as possible, especially, to realize so-called system-on-chips (SOCs). Therefore, to realize a single-ended drive balun directly on CMOS is one of the key innovations for SOCs.

In this part, two single-ended drive baluns operating in the frequency ranges of 800MHz $\sim$ 2.5GHz and 1.5 GHz $\sim$ 4.0 GHz respectively are introduced and studied experimentally. An asymmetrical configuration and tuning capacitors are introduced to further improve the balun performance. In Section 6.2.1, the miniaturized multilayer symmetrical stacked balun (MSSB) configuration is presented. In Section

6.2.2, the realization of the baluns is introduced. In Section 6.2.3 and 6.2.4, two single-ended drive baluns are implemented on a commercial CMOS technology and measured by on-wafer test set. In Section 6.2.5, the parasitic effects of the designed balun are analyzed. A configuration with improved balance by using asymmetrical layout is demonstrated. In Section 6.2.6, loss reduction method by using tuning capacitors is introduced and discussed.

6.2.1 Balun construction and design

A. Symmetrical and stacked configuration

The symmetrical spiral is shown in Figure 6.7 a). Most metal traces of the symmetric spiral is on the same layer except a number of crossovers or via bridges served as the interconnections in the crossover parts. Due to the symmetrical characteristics, the spiral between the two ports and the common node form two inductors L_1 and L_2 with the same inductance. Since the metal traces from the ports to the common node are symmetrical and intervallic, there are the mutual inductances M_{12} and M_{21} (M_{12} is equal to M_{21} for a passive, reciprocal transformer) exist between L_1 and L_2 .

The symmetrical spiral described above are used as the primary spiral of the balun. The same type spiral shown in Figure 6.7 b), which form symmetrical inductors L_3 and L_4 , are used as the secondary spiral. When the secondary spiral is stacked over the primary spiral (i.e. L_1 over L_3 , L_2 over L_4) with the common nodes confronting each other, the multilayer symmetrical stacked balun (MSSB) is formed

by this four port stacked structure, which provide best area efficiency by mutual magnetic enhancements [27]-[28] (refer to the directions of the magnetic flux in the inner side of the spirals and the current directions in the spirals in Figure 6.7).

Figure 6.7 the construction of the MSSB

× denote that the magnetic flux flows in

B. Differential drive transformer

When the MSSB is driven by differential drive signal in port 1 and port 2 of primary spiral, the output signals in port 4 and port 3 of secondary spiral can also be treated as differential drive source. Thus the balun is fully symmetrical in electric and magnetic coupling. This means both the primary spiral and the secondary spiral with the differential drive signal have a higher Q-factor and broader bandwidth than that with the single-ended drive signal [26]. Thus the MSSB shows the widest bandwidth and the lowest insertion loss as the EM result shown in Figure 6.8. The 3-dB bandwidth covers from 0.83 GHz to 14.5 GHz. The phase error is less than ± 1.2 degree, the amplitude error is less than 0.1 dB, and the insertion loss is around -6 dB

in the operating frequency range.

Under the differential drive in input port, the MSSB can be used for ultra-wide band transformer. However, the differential drive signal must be generate firstly. Therefore, the single-ended drive balun, which can convert the single-ended drive signal to the differential drive signal, need to be designed.

Figure 6.8 Simulation results of MSSB under differential drive

6.2.2 Single-ended balun realization

Single-ended drive MSSB for more general utilization is proposed in following. The input signal is connected to port 1 and port 2 is shorted to the ground. However, the primary spiral is no longer fully symmetrical in electromagnetic (EM) characteristics. The symmetrical characteristics of the secondary spiral are also affected. Electrical characteristics of MSSB with single-ended drive source will be discussed in following sections, and this section mainly introduces the balun construction and measurement.

Figure 6.9 Multilayer balun configuration (not to scale)

The single-ended drive MSSB, which is constructed on CSM 0.18 μm CMOS process with six metal layers and one poly layer, is shown in Figure 6.9. The primary symmetrical spiral occupy the fourth metal layer ($M4$) and the third metal layer ($M3$), while the secondary symmetrical spiral is formed by the sixth metal layer ($M6$) and the fifth metal layer ($M5$). CT is connected to the center point of the secondary spiral. ST or port2 is connected to the ground. In order to symmetrically interconnect to $M4$ and $M6$ respectively, the via-bridges or crossovers are built by using the metal traces on $M3$ and $M5$ as well as vias. For the on-wafer measurement, ST and CT are connected with the guard ring [25] or ground pad (G) through the metal trace in the first metal layer ($M1$). The port 1 is input port. Port 3 and port4 in Figure 6.9 a) are corresponding to the output1 and output 2 in Figure 6.9 b) respectively. Two baluns with different turn numbers and operating frequencies of 800MHz~2.5GHz and 1.5GHz~4GHz are implemented using this CMOS process.

In order to measure the performance of the balun, Agilent 8510C network

analyzer, HP8516A S-parameter test set, HP83623A synthesized sweeper and 50A-GSG-100-VP microprobes from CASCADE MICROTECH are used. The calibration data are taken from the stored values in the analyzer, which are made with the standard short-open-load-through measurements on a GGB CS-5 calibration substrate and the same cables and probes. Figure 6.10 illustrates the balun wafer in the chamber system.

Figure 6.10 The balun wafer in the chamber measurement system

6.2.3 Single-ended drive 800 MHz~2.5 GHz balun

Balun A, as in Figure 6.9, has the primary spiral of 5 turns and secondary spiral of 5.5 turns. The widths of the traces and the gaps between adjacent traces are 8 μm and 3 μm respectively. The size of balun A is 270 μm ×270 μm . The guiding ring is connected to the ground.

The measured results are shown in Figure 6.11 and Figure 6.12. Port 3 and port 2 in these figures are corresponding to output 1 and output 2 of Figure 6.9 b) respectively. For the interested frequency range of 0.8 GHz~2.5 GHz, the phase

errors are less than 3.2 degree and the amplitude errors are less than 0.4 dB as shown in Figure 6.11. The measured phase and amplitude characteristics of the single-ended drive balun from 0.05 GHz to 10 GHz are shown in Figure 6.12. Due to the effects of the inter-winding parasitic capacitance and overlay parasitic capacitances, the electric coupling between the primary spiral and the secondary spiral is frequency dependent. Thus, the magnitude and phase characteristics corresponding to the frequency of two output ports are affected. The parasitic effects will be discussed in Section 6.2.5. The die photograph is illustrated in Figure 6.13. The balun has much smaller size than those reported in the literature. It is only about 2% of the single-ended drive balun reported in [15] and [23].

Figure 6.11 Measured results of single-ended drive balun A

Figure 6.12 Measured and model-based results of single-ended drive balun A

Figure 6.13 Die photograph of single-ended drive balun A with guard ring

6.2.4 Single-ended drive 1.5GHz~4GHz balun

Another balun (called balun B), which has a similar configuration as presented in Figure 6.9 and has the primary spiral of 3 turn and secondary spiral of 3.5 turns, is also fabricated on the same wafer as balun A. The size of the balun is $230\mu\text{m} \times 220\mu\text{m}$ and the operating frequency can cover 1.5 GHz~4 GHz. The measurement results of balun B are shown in Figure 6.14 and Figure 6.15. Port 3 and port 2 in these figures are corresponding to output 1 and output 2 of Figure 6.9 b) respectively. The

measured characteristics from 1.5 GHz to 4 GHz are shown in Figure 6.14. As denoted by the marks, the amplitude and phase imbalances are less than 0.5 dB and 2.0 degrees respectively. The measured amplitudes and phases characteristics from 0.05 GHz to 10 GHz are illustrated in Figure 6.15.

Figure 6.14 Measured results of single-ended drive balun B

Figure 6.15 Measured and model-based results of single-ended drive balun B

Both of the baluns (balun A and balun B) have the same configuration, the same width of the metal trace as well as the same width of the gaps. The only difference is the turn numbers. By comparing Figure 6.11 with Figure 6.14, it can be seen that both the magnitudes and the phase delays are different. For Figure 6.11, the magnitude difference of one dB between S_{3I} and S_{2I} lies in 3.73 GHz, while for Figure 6.14, the magnitude difference of one dB between S_{3I} and S_{2I} is located in 7 GHz. These differences are mainly due to the reduced turn numbers, which can lower the inductance as well as the parasitic capacitances, and can increase the self-resonant frequency. The operating frequency range can be further shifted up with the decrease of the turn number. Thus, the balun configuration can be used for even higher operating frequency range. However, the loss in the low operating frequency range will increase with reduced turn number due to the mismatch at the low frequency range. These parasitic effects will be discussed by equivalent circuit model in following sections.

6.2.5 Parasitic effects analysis

Since an ideal transformer balun is dominated by the magnetic coupling, the single-ended drive function model used in [28] is helpful to understand the magnetic coupling relationships among the inductors in the primary and the secondary spirals. However, the parasitic effects such as parasitic capacitances, resistances and substrate losses can deteriorate the performance of the baluns as demonstrated in the measured frequency point, where the magnitude difference between the two output signals is

one decibel. Figure 6.16 shows the equivalent-circuit model. The port 1 represents input port in Figure 6.9 b). Port 2 and port 3 represent the output2 and output 1 of Figure 6.9 b) respectively. As denoted by L_1 , L_2 , L_3 and L_4 as well as the mutual magnetic coupling M_{ij} ($i=1,2,3,4$, $j=1,2,3,4$ $i \neq j$, and $M_{ij} = M_{ji}$), the MSSB function model in [28] is embedded into the equivalent-circuit model to represent the transformer as shown in Figure 6.16. R_{si} corresponding to L_i ($i=1, 2, 3, 4$) represent the ohmic losses of the inductors. The shunt resistor R_{sub} and the shunt capacitor C_{sub} to the ground as well as C_{ox} represent dissipations and parasitic capacitances in the substrate and oxide layer. Three capacitors C_{s1} , C_{s2} and C_{s3} , which are respectively connected between each port of three ports and the ground, represent the capacitances from the ports to the ground. C_{p1} and C_{p2} represent the sum of the stack and fringe capacitances between L_1 and L_3 and between L_2 and L_4 respectively. C_g represents the center tap of the primary spiral to ground capacitance. C_{p3} represents the capacitance between the port 2 and port 3, while C_{p4} models the parasitic capacitance between port1 and port2.

Figure 6.16 Single-ended drive Multilayer balun model

The comparison of the model-based simulation results and the EM simulation results for balun A and balun B is shown in Figure 6.12 and Figure 6.15 respectively. Good agreements between the measured and modeled S -parameters are achieved below the first self-resonant frequency. The equivalent circuit model can only represent the baluns below the first self-resonant frequency. When the ports to ground parasitic capacitances (i.e. C_{s1} , C_{s2} and C_{s3}), C_g , C_{p4} , C_{p1} and C_{p2} are set as zero, both amplitude and the phase of the signals in two output ports of the balun remain balanced (i.e. the two outputs are equal in amplitudes and 180 degrees out of the phase). It means that changing the values of L_i and the mutual magnetic coupling coefficient K_{ij} [28] can only affect the absolute amplitudes and phase delays. However, when the above conditions are not matched, the balun characteristics will be affected in higher operating frequency range where the electric coupling effects can not be neglected. The voltage difference between port1 and port3 is about three times as large as that between port2 and the ground. The electric couplings from port1 to port3 as well as from port2 to the ground increase with the frequency. Thus the absolute voltage difference between port 2 and port3 is increased. At the same time, the parasitic capacitances of C_{s1} , C_{s2} , C_{s3} and C_{p3} , which are smaller than C_{p1} or C_{p2} , can also affect the balance characteristics. In Figure 6.11 and Figure 6.14, the absolute amplitude of S_{31} is larger than that of S_{21} in the higher operating frequency range. From investigating the model-based results, it is found that when the voltages at the three ports are fixed, decreasing the capacitance of C_{p1} can reduce the electric coupling from port1 to port3 and thus decrease the magnitude of S_{31} . At the same time,

the decreasing capacitance of C_{p2} can also reduce the electric coupling from port2 to the ground and thus increase the magnitude of S_{21} . From the above analysis, it is clear that reducing the capacitances of C_{p1} and C_{p2} can improve the amplitude balance of the balun. There are two ways to reduce the capacitances of C_{p1} and C_{p2} . The first method is to increase the distance between the primary spiral and the secondary spiral. But, the increased distance can reduce the magnetic coupling between the primary spiral and the secondary spiral. Further, when the metal traces of the spirals are closing to the silicon substrate, more substrate loss will be generated. Thus, the insertion loss of the balun will be increased.

The second method is to decrease the width of the stacked metal traces. However, the metal loss will be increased and magnetic coupling will be decreased due to the reduced width of the metal traces. Here, we demonstrate an asymmetrical configuration, where inductor 1 and inductor 2 of the primary spiral are on $M3$ layer and $M4$ layer respectively. Both inductor 3 and inductor 4 of the secondary spiral are still on the same metal layer ($M6$) except the crossovers of via bridges. Both the magnetic and the electric couplings from port1 to port3 of the asymmetrical balun configuration are reduced as compared to the symmetrical configuration of single-ended drive MSSB in Figure 6.9. However, the electric coupling from port3 to the ground is almost unaffected. Thus the amplitude balance can be improved. This analysis is verified by full-wave EM simulation. The phase and the amplitude characteristics of a symmetrical and asymmetrical configuration with the primary spiral of 3 turns and secondary spiral of 3.5 turn are compared in Figure 6.17. For the

asymmetrical configuration, the amplitude of S_{21} is reduced, while the amplitude of S_{31} is almost the same as that of the symmetrical configuration. Below 4 GHz, the amplitude and phase imbalance are less than 0.29 dB and 1.0 degree (or 0.0175 rad) respectively. The one-dB magnitude difference between S_{31} and S_{21} is located in 8.1 GHz. Thus the balance characteristics of the balun are well improved.

Figure 6.17 Comparison of the symmetrical and asymmetrical baluns

6.2.6 Balun loss reduction

Generally speaking, for the ideal balun with magnetic transformer, the EM energy in the interested RF/Microwave frequency range is mainly stored and transferred in the magnetic energy. For the LC tank, the EM energy can be stored and transferred in both the inductor with the magnetic energy and the capacitor with electric energy simultaneously. Since the quality factor of the metal-isolator-metal (MIM) capacitor on CMOS is much larger than that of the inductor on CMOS, the capacitors can be used to improve the performance of the balun in the following.

Figure 6.18 Schematic of single-ended drive balun with tuning capacitors

As the schematic illustrated in Figure 6.18, the impedances of two loads Z_{L1} and Z_{L2} and the source impedance Z_g are set to 50Ω . A shunt capacitor C_1 at input port and a parallel capacitor C_2 between two output ports are used. Firstly, balun A is used in the schematic of Figure 6.18. It is demonstrated that the capacitors can reduce the transmission loss up to 1.5 dB in the range of 800 MHz~1.5 GHz. The phase delays from the input port to two output ports are shifted, while the phase errors or phase imbalance can be improved when two proper capacitances are set (see the example shown in Figure 6.19). The performance in low frequency range can be further improved by setting $C_1=1.27$ pF and $C_2=1.52$ pF. The insertion loss in the range of 850 MHz ~ 1.05 GHz can be reduced to less than 5.4 dB. The loss reduction comes from the improved ports matching as well as smaller loss of the added capacitances. In the interested frequency range, the balance characteristics of the phases and the amplitudes of these two output signals are not deteriorated by the tuning capacitors. However, this loss reduction relies on the resonance of the LC tankers, which are formed by inductances L_i , parasitic capacitances and the added capacitances in Figure 6.18. Hence, the reduced loss can be obtained only in narrow band range. This capacitor tuning method is useful to improve the performance of the single-ended

drive MSSB.

Figure 6.19 Balun characteristics under two different tuning conditions

As a further example, the asymmetrical single balun in Figure 6.17 is also used in the schematic of Figure 6.18. As demonstrated in Figure 6.20, when $C_1=1.1$ pF and $C_2=1.2$ pF, both the amplitude and phase characteristics are improved in the low operating frequency range. Under this condition, the insertion loss is only 4.7 dB (Balun loss for each channel is only about -1.7 dB, since the ideal balun has insertion loss of 3 dB for each channel). The phase difference between two ports is improved and is quite closed to π (or 180 degrees) from 800 MHz to 3 GHz. By comparing the characteristics of the tuning balun in Figure 6.19 and Figure 6.20, it can be found that the tuning frequency range is broadened and the insertion loss at the resonant frequency is reduced due to the reduced turn number of the spirals. The reduced turn number can reduce the self-inductances and mutual inductances simultaneously. Thus the resistances or losses of the spirals are reduced. On the other hand, to keep the

same operating frequency as that for the balun without reducing turn number, the capacitances must be increased. However, the parasitic loss or resistance of the capacitances is much less than that of the inductances on the CMOS process. Thus the loss of the single-ended drive balun can be greatly reduced up to 4.8 dB by reducing the turn number and introducing the capacitors with reasonable capacitances. These characteristics will be helpful to further reduce the loss for narrow band operation of the single-ended drive balun.

Figure 6.20 Comparison of the asymmetrical balun with and without tuning

6.3 Conclusion

In this chapter, the miniaturized DMML ring resonator and the single-ended drive MSSBs are proposed and implemented for the first time. The performance degeneration of the DMML ring resonator due to the high loss P-silicon substrate is shown experimentally. The technique of mutual magnetic coupling enhancement is

employed in the design of the MSSB. The designed balun A, which is the smallest balun in the world, can cover all the commercial mobile bands such as AMPS, GSM900/1800, DECT, PHS, PDC, GPS, and 2.45 GHz ISM-band such as IEEE 802.11b/g as well as the Bluetooth. The designed balun B can cover frequency range of 1.5 GHz~4 GHz. The equivalent-circuit model is used to analyze the parasitic characteristics. The asymmetrical configuration and capacitor tuning method are proposed based on the model-based analysis to further improve the balance and the loss characteristics of the balun.

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CHAPTER 7

SUMMARY AND RECOMMENDATIONS

7.1 Summary

In this dissertation, RF/Microwave passive circuits including resonators, filters and baluns are extensively investigated both theoretically and experimentally. Several novel approaches are used to design these passive circuits with miniaturized size and good performance. The passive circuits are successfully implemented on different fabrication technologies such as PCB, silicon micromachining, thin-film as well as commercial CMOS technologies.

Beside the implemented microwave circuits, passive structures and their characteristics are extensively investigated. Different transmission lines including MS line, CBCPW, CBCPS and different microwave resonator circuits including ring resonator, patch resonator, modified quarter-wavelength resonator, etc. are also discussed. New SEMCP filter topology, EM-based parameter extraction, magnetic coupling enhancement and cell cascading with phase compensation are proposed and analyzed. The general SEMCP filter topology is proposed to achieve quasi-elliptical filter response with non-cross-coupled resonators. The implemented filter based on proposed filter topology has much compact size and high performance. The magnetic coupling enhancement, model-based parasitic effect analysis and capacitor tuning approaches are introduced to the single-ended drive balun design. A world smallest balun with good performance is achieved on standard CMOS technology. In addition,

the modeling of microwave structure by using lumped circuit model, which is important for designing and understanding of microwave circuit, is studied. Both the scalable model and non-scalable model are investigated. The suppression and shielding of leakage power, which is quite important for packaging design, are also considered by both analysis and experiment. The content of this dissertation should be useful for the designer in the related area.

Major contributions reported in this thesis include:

1) A general filter topology, configuration and design method based on SEMCP are proposed. Based on the SEMCP filter topology, second-order and high-order SEMCP filters with multiple finite transmission zero points are implemented on standard single layer PCB board. The characteristics of the electric dominant and magnetic dominant SEMCP filters are compared and the advantages such as compact size and low loss are demonstrated. Based on the proposed method and topology, the elliptical response, which can only be achieved by using cross-coupled topology, can be achieved directly by using the cascaded resonators with SEMCP. The method can be applied to different types of transmission lines such as stripline, coplanar waveguide and suspended stripline to achieve improved performance as well as reduction in size.

2) The novel filter and diplexer with compact size and multiple finite transmission zero points using modified open-ground structure and source-load coupling scheme are analyzed and implemented.

3) A novel spurious suppression method for the quarter-wavelength resonator and

filter is proposed to suppress the spurious frequency at $3f_0$ (where f_0 is the operating frequency of the filter). As an example, a filter with wide stopband bandwidth up to $5f_0$ and large stopband rejection has been demonstrated. The method can be used in input/output (I/O) ports as well as inter-stage coupling of quarter-wavelength filters (like interdigital and combline filter) to achieve wider stopband bandwidth.

4) For the first time, the gradual periodic loading of conductor backed coplanar waveguide (CBCPW) transmission lines and filters are investigated with consideration of the power leakage. The leakage characteristics of the CBCPW are discussed based on the cell parameter-extraction, cell cascading and experiment. A novel method of cell cascading with phase compensation is proposed. It is used to implement the miniaturized CBCPW lines and filter by using the thin-film technology.

5) A ground-ring guarded patch resonator, which has lower radiation loss and substrate leakage than the traditional patch resonator, is proposed and implemented on silicon wafer for the first time. A ground-ring guarded patch resonator with ground shunt is used to realize the filter with size reduction up to 70%. All these designs are fabricated on silicon substrate using SiDeox (silicon deep etching and oxidation) and through hole plating technology in our laboratory.

6) Based on the mutual magnetic coupling enhancement technique, a novel ultra-wide bandwidth differential drive transformer configuration and single-ended drive multilayer symmetrical stacked balun configuration are proposed and implemented using CSM (Chartered Semiconductor Manufacturing) 0.18 μ m CMOS

technology. A balun, with frequency range of 800MHz ~2.5GHz, can cover all the commercial mobile bands such as AMPS, GSM900/1800, DECT, PHS, PDC, GPS, and 2.45GHz ISM-band such as IEEE 802.11b/g as well as the Bluetooth. Another miniaturized balun can cover the frequency range of 1.5GHz~4GHz. The asymmetrical configuration and resonant tuning method are illustrated to improve the balance and the loss characteristics. Although the proposed baluns are only demonstrated in commercial CMOS technology, it is also applicable to other technologies, such as GaAs.

7.2 Recommendations

1) The novel SEMCP filter topology in Chapter 3 can be used to implement lumped element filters or other transmission line filters such as the CPW filter and waveguide filter. The OG and SEMCP configurations can be modified to form the cascade quadruplet (CQ) and cascade trisection (CT) configurations with high selectivity, compact size and more transmission ZPs.

2) The method of cell cascading with phase compensation for CBCPW periodic loading transmission lines in Chapter 4 can be extended to cell cascading with both amplitude and phase compensation by using the transmission matrix cascading method and the full-wave FDTD analysis. Thus the cascaded linear error can be compensated, and computation should also be reduced.

3) The equivalent-circuit model proposed in Chapter 6 can be extended to a scalable model. The asymmetrical configuration and resonant tuning method to

improve the balance and the loss characteristics, which are illustrated in the simulations, can be realized by using MIM capacitors. The balun can be used to design different active circuits such as the push-pull power amplifier, differential mixer and VCO.

4) The operating frequency of the patch resonator with ground shunt in Chapter 5 is calculated by full-wave analysis. Actually, the general closed form expression for operating frequency and the quality factor could be also extracted. Thus the initial design can be easily performed without using the full-wave EM simulation.

5) The U-shape OG filter and diplexer with compact size and multiple ZPs using OG structure and the source-load coupling scheme shown in Chapter 3 can be implemented by using multi-layer PCB technology or LTCC technology, and further size reduction can be expected. By adding the tunable MEMS capacitors or varicaps at the open-ends of the OG filter and at the source-load coupling gap of the OG filter, the filter with both tunable operating frequency and bandwidth can be achieved. The spurious suppression method at I/O ports in Chapter 3 can be combined with PBG structure to suppress the spurious frequency and thus extend the stopband bandwidth.

Author's Information

Mr. Kaixue Ma received the B.Sc. and M.Eng. degree in electronics engineering from Northwestern Polytechnical University (NWPUE), P.R. China, in 1997 and 2001 respectively. From 1997 to 2002, he was with 504th Institute of China Academy of Space Technology (CAST), where he became a deputy director of MM-wave group and focused on the design of space borne RF, Microwave and MM-wave active components and subsystem for satellite transponder system and VSAT base station. He is currently working toward his Ph.D degree at the School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore. His research interests include the design and modeling of passive RF IC circuits on CMOS, MEMS and PCB processes.

Research related to this thesis has resulted in the following publications:

Patent:

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