Novel Structures and Thin Film Techniques for Reconfigurable RF Technologies With Improved Signal Integrity

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NOVEL STRUCTURES AND THIN FILM TECHNIQUES FOR RECONFIGURABLE RF TECHNOLOGIES WITH IMPROVED SIGNAL INTEGRITY

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DEDICATION

To my beloved family
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ABSTRACT

Over the past several decades, wireless communications have been tremendously developed with multiple frequency bands, standards, and functions to provide more convenient communication services. Nowadays, commercial wireless devices, especially customer premises equipment, are required for smaller size, higher integration, and higher data rates to meet the ever-increasing demands of communications. The transceiver front ends need to provide high-speed and high-quality communication services without a substantial increase in cost and size. This dissertation aims to develop miniaturized, multifunctional, and reconfigurable RF technologies and to signal integrity improvement solutions for highly integrated systems with high data transfer rates.

A promising way to shrink the communication systems is to replace the repetitive and cumbersome discrete RF passive devices with multiband and multifunctional RF components. A filtering structure is first developed in this dissertation to provide a dual-band filtering response with high selectivity. Multiple transmission poles are generated by utilizing half-wavelength and open stub loaded resonators. In addition, a novel coupling scheme is applied to the design to generate multiple transmission zeros, leading to highly selective passband and good out-of-band rejection. This filtering structure is then applied to design a switch-controlled reconfigurable dual-band bandpass filter and a dual-functional filtering balun. The excellent performances of the implemented filter and
filtering balun show great superiority and demonstrate the design efficacy of the proposed filtering structure.

Besides designing multifunctional or reconfigurable RF components in specific configurations, this dissertation investigates a more flexible solution to facilitate the reconfigurability and miniaturization of the integrated systems. Designing tunable and miniaturized RF components on engineered substrate enabled with ferromagnetic thin films is fully studied in the dissertation. By selectively patterning the ferromagnetic thin films, the engineered substrate provides increased and tunable effective permeability with limited losses. In order to achieve higher permeability, more or thicker ferromagnetic thin film layers are required. The performance of the RF components is dependent on the properties of the engineered substrate. It is critical to develop an accurate model which quickly optimizes and determines the configuration of the ferromagnetic thin film patterns in the engineered substrate for a desired permeability. In this dissertation, a multilayer substrate model is created by thoroughly exploring the impact of different factors on the performance of the engineered substrate, such as pattern dimensions, the number of film layers, film thickness, filling density, etc. To demonstrate the design efficacy of the developed model, a miniaturized and tunable frequency selective surface (FSS) and a performance-enhanced antenna are implemented on the optimized engineered substrate.

With the previous designs and techniques, a highly integrated system with multiple frequency bands and multiple functions is achieved. However, a highly integrated system suffers the signal integrity issue due to the far-end crosstalk in high-density traces, especially as the data transfer rates get higher and higher. Adding tabs between the coupled lines is an existing good way to reduce crosstalk. In this dissertation, thin film techniques
are applied to further reduce the far-end crosstalk and improve the signal integrity of the highly integrated system. The far-end crosstalk of coupled transmission lines and the tabbed routing structure is thoroughly analyzed, and closed-form formulas are derived to characterize and optimize the performance. Based on the theoretical analysis, practical solutions are developed to further reduce the crosstalk between interconnects, including the application of ferromagnetic thin films and ferroelectric thin films and new interconnect structure with non-uniform signal line tab thickness.
# TABLE OF CONTENTS

Dedication ........................................................................................................ iii
Acknowledgements ................................................................................................. iv
Abstract ..................................................................................................................... vi
Preface ...................................................................................................................... viii
List of Tables ........................................................................................................... x
List of Figures .......................................................................................................... xi
List of Symbols ......................................................................................................... x
List of Abbreviations ............................................................................................... xvi

Chapter 1 Introduction ........................................................................................... 1
  1.1 Motivation and Background .......................................................................... 1
  1.2 Chapter Outline ............................................................................................. 16

Chapter 2 A Dual-Band Filtering Structure for Highly Selective Reconfigurable Bandpass Filter and Filtering Balun .......................................................... 19
  2.1 Introduction .................................................................................................. 19
  2.2 Basic Dual-Band Filtering Structure ............................................................. 22
  2.3 Switchable Dual-Band Bandpass Filter ......................................................... 26
  2.4 Dual-Band Bandpass Filtering Balun ............................................................. 33
  2.5 Conclusion .................................................................................................. 38

Chapter 3 Modeling of Engineered Substrate enabled with ferromagnetic thin films ...... 40
  3.1 Introduction .................................................................................................. 40
3.2 Characterization of Permalloy Enabled Engineered Substrate..........................46
3.3 Model Verification and Design Examples ...................................................58
3.4 Conclusion..................................................................................................64
Chapter 4 High-Performance Interconnects with Improved Signal Integrity........66
  4.1 Introduction ..............................................................................................66
  4.2 Far-End Crosstalk (FEXT) Analysis ..........................................................68
  4.3 Methodologies for FEXT Reduction ..........................................................73
  4.4 Fabrication and Measurement .................................................................79
  4.5 Conclusion ..............................................................................................86
Chapter 5 Summary and Future Work ...............................................................87
  5.1 Summary of Contributions .......................................................................87
  5.2 Future Work ............................................................................................90
References .......................................................................................................92
Appendix A: Multiport VNA Measurement .......................................................101
LIST OF TABLES

Table 2.1 Four-state filtering responses summary ........................................................... 27
Table 2.2 Parameter values used in the switchable BPF ..................................................... 28
Table 2.3 Comparison with similar references ................................................................. 32
Table 2.4 Parameter values used in the filtering balun ..................................................... 35
Table 2.5 Comparison with similar references ................................................................. 37
Table 3.1 Effective permeability calculated by the proposed model ................................. 58
Table 3.2 Comparison of antenna performance ............................................................... 64
Table 4.1 Comparison of capacitance and inductance ....................................................... 72
Table 4.2 Comparison of capacitance and inductance of the tabbed routing structure ..... 75
Table 4.3 Calculated capacitance, inductance and $|k_f|$ for different cases ....................... 79
Table 4.4 Specific measured FEXT ($S_{41}$) values versus frequencies for different structures (dB) ......................................................................................................................... 83
LIST OF FIGURES

Figure 1.1 Illustration of the wireless communication systems supporting multi-functional services with multiple frequency bands and wireless standards........................................1

Figure 1.2 Sub-6 GHz RF front-end structure for a mobile device. ......................................3

Figure 1.3 Diagram of a reconfigurable, multifunctional and miniaturized system with tunable and improved RF components. .................................................................4

Figure 1.4 Miniaturized systems enable smaller mobile devices while the crosstalk between high-density traces degrades the signal integrity.................................5

Figure 1.5 Examples of multiband filters implemented in different technologies: (a) triple-band filter by waveguide [10], (b) Multiband filter by substrate integrated waveguide (SIW) [11], (c) Dual-band filter by acoustic-wave resonator [12], (d) Multi-passband filter by microstrip line [13].................................................................6

Figure 1.6 Examples of tunable RF components by different tuning technologies: (a) mechanical tuning system for tunable filters [18], (b) tunable filter by varactors [19], (c) MEMS-based tunable filter [20], (d) tunable capacitor by ferroelectric material [21], (d) tunable resonator b ferromagnetic materials [22].................................................10

Figure 1.7 Technologies for crosstalk reduction: (a) covering dielectric film [32], (b) inserting guard trace [33], (c) introducing groove (groove thickness is described by $d$) [36], (d) using decoupling capacitance [37], (e) adding tabs [38].................................14

Figure 2.1 (a) Typical configuration of a front-end module. Simplified module enabled with (b) dual-band switchable filter and (c) dual-band balun filter...............................20
Figure 2.2 (a) Simplified open-stub loaded resonator and its equivalent circuit under (b) odd-mode, and (c) even-mode .................................................................22
Figure 2.3 (a) Basic dual-band filtering structure, (b) coupling scheme of the filtering structure..................................................................................................................24
Figure 2.4 Architecture of the switchable dual-band BPF........................................26
Figure 2.5 Signal suppression technologies utilized in the filter: (a) one-end-grounded coupled line, (b) stepped-impedance open stub.................................................27
Figure 2.6 Photograph of the fabricated PIN diodes controlled reconfigurable BPF, and biasing circuit of diode (insert) .................................................................................................28
Figure 2.7 Simulated and measured S-parameters of the designed switchable BPF: (a) both passbands ON, (b) both passbands OFF, (c) low-frequency passband ON, (d) high-frequency passband ON .........................................................................................30
Figure 2.8 (a) 3D view of the dual-band balun BPF. (b) Schematic layout of the top layer .................................................................................................................................33
Figure 2.9 Photograph of the fabricated balun BPF: (a) front view, (b) back view ........34
Figure 2.10 Simulated and measured results of the balun BPF: (a) S-parameters, (b) phase and magnitude imbalances of each passband............................................................................36
Figure 3.1 Engineered substrate enabled by patterned ferromagnetic thin films .........41
Figure 3.2 (a) Unmagnetized-ferromagnetic, (b) magnetized-ferromagnetic ............42
Figure 3.3 (a) Hysteresis loop and (b) electrical tuning method of ferromagnetic materials .................................................................................................................................43
Figure 3.4 Microstrip line model with ferromagnetic thin films ................................47
Figure 3.5 The characteristic impedance of a microstrip line converges at a certain substrate width, which is defined as the effective width where ferromagnetic thin films have significant effects on the extracted inductance. (b) Fitting surface of the effective width ($W_{conv}$) by least squares method

Figure 3.6 The real model of a microstrip line with the engineered substrate is considered as the equivalent model with effective permeability ($\mu_{eff}$) when both models have the same extracted inductance

Figure 3.7 (a) Simulated results of microstrip lines with different dimensions. (b) Comparison of the results generated by HFSS simulation and the fitting formula

Figure 3.8 Extracted inductance of the microstrip line model under different pattern sizes and orientations

Figure 3.9 Derivation of correction factors in different scenarios: (a) vertical position, (b) film thickness, (c) number of film layers, and (d) planar filling density

Figure 3.10 A square ring based FSS on the engineered substrate

Figure 3.11 (a) Results comparison of the FSS with and without the engineered substrate. (b) Performance of the miniaturized FSS under different DC biasing conditions

Figure 3.12 Patch antenna on the optimized engineered substrate

Figure 3.13 (a) Performance comparison of miniaturized antenna (design 1) on the engineered substrate and original antenna on a regular dielectric substrate. (b) Performance comparison of the original antenna and design 2 with the same miniaturization factor

Figure 4.1 (a) Equivalent circuit of the coupled lossless microstrip lines. (b) Capacitive components of coupled microstrip lines with arbitrary line widths
Figure 4.2 (a) Tabbed routing structure. (b) Equivalent division of the unit segment. (c) Additional fringing capacitance to consider when calculating mutual capacitance

Figure 4.3 Three improved structures to further reduce the FEXT: (a) 3D tabbed routing structure with thicker tabs, (b) tabbed routing structure with dielectric material between tabs, (c) tabbed routing structure with covered magnetic material

Figure 4.4 Comparison of the simulated FEXT ($S_{41}$) of various interconnects

Figure 4.5 (a) Specially designed circuit for measurements by the RF probe station. (b) Microstrip line to CPW transition for probe measurements

Figure 4.6 Schematic diagram of the fabrication process flow. Steps (3a), (3b), and (3c) were performed on separate wafers to independently assess the impact on device performance

Figure 4.7 Photographs of the fabricated devices: (a) coupled lines, (b) tabbed routing structure, (c) tabbed routing structure with thicker tabs, (d) tabbed routing structure with BST, (e) tabbed routing structure with permalloy

Figure 4.8 Comparison of the measured and simulated (a) FEXT ($S_{41}$) and (b) insertion loss($S_{21}$)

Figure 4.9 Circuits for eye diagram simulations in ADS based on the measured SNP files

Figure 4.10 Eye diagrams of (a) coupled lines, (b) tabbed routing structure, (c) tabbed routing structure with thicker tabs, (d) tabbed routing structure with BST, (e) tabbed routing structure with permalloy
LIST OF SYMBOLS

\( \varepsilon \)  Permittivity
\( \mu \)  Permeability
\( \delta \)  Skin depth
\( k_f \)  Far-end coupling coefficient
\( L \)  Inductance
\( C \)  Capacitance
\( Ti \)  Titanium
\( Au \)  Gold
\( S \)  Scattering parameter
LIST OF ABBREVIATIONS

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>ADS</td>
<td>Advanced Design System</td>
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<tr>
<td>Ant</td>
<td>Antenna</td>
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<td>AT</td>
<td>Aperture Tunner</td>
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<tr>
<td>BPF</td>
<td>Bandpass Filter</td>
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<tr>
<td>BST</td>
<td>Barium Strontium Titanate</td>
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<tr>
<td>CPW</td>
<td>Coplanar Waveguide</td>
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<tr>
<td>DC</td>
<td>Direct Current</td>
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<tr>
<td>DUT</td>
<td>Device Under Test</td>
</tr>
<tr>
<td>DP3T</td>
<td>Double Pole Trip Throw</td>
</tr>
<tr>
<td>DRX</td>
<td>Discontinuous Reception</td>
</tr>
<tr>
<td>DSPSL</td>
<td>Doubled-Sided Parallel-Strip Line</td>
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<tr>
<td>EM</td>
<td>Electromagnetic</td>
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<tr>
<td>ET</td>
<td>Envelope Tracking</td>
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<tr>
<td>FEXT</td>
<td>Far-end Crosstalk</td>
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<tr>
<td>FMR</td>
<td>Ferromagnetic Resonance</td>
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<tr>
<td>FSS</td>
<td>Frequency Selective Surface</td>
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<tr>
<td>GSG</td>
<td>Ground-Signal-Ground</td>
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<tr>
<td>GSM</td>
<td>Global System for Mobile Communications</td>
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<tr>
<td>HFSS</td>
<td>High Frequency Structure Simulator</td>
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<tr>
<td>HN-LTE</td>
<td>High-Band Long Term Evolution</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate Frequency</td>
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IPD ............................................................... Integrated Passive Device
IT ........................................................................ Impedance Tuner
LB-LTE ................................................................ Low-Band Long Term Evolution
LNA .................................................................... Low-Noise Amplifier
MB-LTE ................................................................ Mid-Band Long Term Evolution
MEMS ................................................................ Microelectromechanical Systems
NEXT ....................................................................... Near-end Crosstalk
PA ........................................................................ Power Amplifier
PCB ........................................................................ Printed Circuit Board
PIN ................................................................ Positive-Intrinsic-Negative
PMIC ................................................................ Power Management Integrated Circuit
PRBS .................................................................... Pseudorandom Binary Sequence
Py ....................................................................... Permalloy
RAC ................................................................ Radio Corporation of America
RF ........................................................................ Radio Frequency
Rx .......................................................................... Receiver
SF ...................................................................... Slowing Factor
SIW .................................................................. Substrate Integrated Waveguide
SOLT ................................................................... Short-Open-Load-Through
SMA ................................................................ Subminiature Version A
TOM .................................................................. Through-Open-Match
TRL .................................................................. Through-Reflection-Line
Trxv ..................................................................... Transceiver
Tx .......................................................................... Transmitter
UHB-LTE .......................................................... Ultra-High-Band Long Term Evolution
UWB ................................................................. Ultra-Wide Band
VNA ........................................................................... Vector Network Analyzer
Xtlk ................................................................. Crosstalk
CHAPTER 1
INTRODUCTION

1.1 Motivation and Background

The rapidly growing wireless communication market has undergone tremendous changes in the requirements and capabilities of radios to support wireless connections [1]. For example, the 850 MHz and 1900 MHz frequency spectrums were used to support the Global System for Mobile Communications (GSM) networks. The modulation method changed with the arrival of the second generation (2G) networks, but mostly the same

![Diagram of wireless communication systems]

Figure 1.1 Illustration of the wireless communication systems supporting multi-functional services with multiple frequency bands and wireless standards.
frequency spectrum was used. New frequency bands and a spectrum of nearly 2100 MHz were adopted with the arrival of the 3G network. As the fourth generation (4G) was deployed, more frequency bands and spectrums were introduced, ranging from 600 MHz to 2.5 GHz [2]. Nowadays, wireless communication systems supporting high-speed internet connectivity with higher data rates are highly demanded, and they play critical roles in the economic development and digitization of society and the world [3]. To support the continuous evolution, the fifth generation (5G) mobile networks have been widely developed [4]. The frequency band of sub-6 GHz ranging between 450 MHz to 6 GHz and millimeter wave (mm-Wave) spectrum ranging from 24 GHz to 52 GHz are employed to provide multifunctional services, enabling 5G to be an ideal solution for urban areas due to the very large bandwidth [5]. The trends of driving new bands and modes of operation will continue in the future 6G systems. As shown in Figure 1.1, wireless communication systems with multiple frequency bands, multi-function, high integration, and high data rates have been widely applied to modern life, keeping everyone and any devices connected at anytime and anywhere.

Nowadays, ultra-reliable and low latency 5G wireless communication systems are required to support more than 50 frequency bands and multiple standards. However, each standard has its own unique characteristics, constraints and specific challenges, it is highly desirable for mobile devices to roam seamlessly across geographic boundaries, accommodating different protocols and frequencies of use as required. This requires the front-end integration of multifunctional modules within a compact area within a mobile device. To achieve such ability, low cost and high linearity radio frequency (RF) front ends with multiple filters are required to access various antennas for 5G mobile devices as
depicted in Figure 1.2 [6]. The support of different frequency bands and wireless standards is achieved by employing parallel integration of separate RF front ends. A number of circuits are duplicated for each band, with the associated penalties of increased size, weight and cost. In addition, a large die area is required for multiple off-chip passive components, such as couplers, filters, diplexers, antennas, etc. However, the board area is very limited due to the device convergence trend. Despite many years of research, there is a technological barrier for further integration to achieve miniaturized multiband communication systems. The ensuing high system complexity coupled with stringent
requirements of low-cost and low power imposes unique challenges to the RF system designers.

To tackle these challenges and enable the continuation of the market requirements of transceivers for multi-standard/multi-function, the well-known Software Defined Radio (SDR) [7] and Cognitive Radio (CR) [8] concepts have been developed in the system level. In general, SDR and CR systems are exploited to sense the frequency spectrum and adopt their operating parameters to the requested bands. This typically requires the ability to reconfigure the radio’s hardware and particularly the RF front-end, over a wide frequency range. Therefore, high-performance, multifunctional, and widely tunable RF components are crucial for the implementation of a fully reconfigurable and miniaturized system. Figure 1.3 portrays a diagram of a simplified and miniaturized system enabled with multiband, multifunctional and tunable RF components. Well-developed RF components with
frequency tunability, muti-function, multi-band and good miniaturization can support different operating frequency bands adapt to the system, resulting in a miniaturized, frequency-agile, and multifunctional system.

With the development of miniaturized and multifunctional systems, mobile devices are getting smaller and smaller, as shown in Figure 1.4. Nevertheless, the issue of signal integrity arises due to the crosstalk between high-density traces in the miniaturized systems. In inhomogeneous environments, the closely placed traces generated strong electric and magnetic coupling, resulting in the crosstalk noise interference. Such interference is generally represented in terms of near-end crosstalk (NEXT) and far-end crosstalk (FEXT). The NEXT will keep as a constant, however, the FEXT will keep increasing as the trace length increases. Therefore, it is very necessary to apply technologies to reduce the FEXT such that the system signal integrity is improved. In
general, the main purposes of this dissertation are to provide multi-functional and reconfigurable RF technologies to enable miniaturized systems and solve the FEXT issue of the highly integrated system. The general overview and comparison of different technologies and the description of the proposed designs and methodologies are delivered in the following sections.

1.1.2 Multiband and Multifunctional RF Components

An important approach to miniaturizing the communication systems is to replace the repetitive and cumbersome RF passive devices with multiband and multifunctional RF

![Figure 1.5 Examples of multiband filters implemented in different technologies: (a) triple-band filter by waveguide [10], (b) Multiband filter by substrate integrated waveguide (SIW) [11], (c) Dual-band filter by acoustic-wave resonator [12], (d) Multi-passband filter by microstrip line [13].](image-url)
components. Multiband cognitive radio has been widely recognized as one of the prominent solutions to tackle the spectrum scarcity with significantly enhanced throughput and better channel maintenance of communication networks [9]. Multiband and multifunctional components such as bandpass filters (BPFs) are of great importance in improving system spectral efficiency, system space-consuming, and multi-function capability.

As shown in Figure 1.5, multiband filters have attracted great attention with a large number of circuit topologies implemented in different technologies such as waveguide [10], substrate integrated waveguide (SIW) [11], acoustic-wave resonator [12] or microstrip line [13]. Waveguide filters are RF filters that use hollow, conductive metal structures to carry RF signal. The capability of carrying high-frequency signals with very low loss makes the waveguide a practical solution in microwave and millimeter-wave bandwidths. However, the bulky metal structure of a waveguide filter always leads to high weight and high cost. It is also not suitable for operations at lower frequencies due to the increased dimensions. To overcome this, substrate integrated waveguide (SIW) is developed on the printed circuit board (PCB) to replace the original waveguide to a certain extent. The SIW is usually formed by adding a top metal over the ground plane and caging the structure with rows of plated vias on either side. The vias in the substrate and the top and bottom metal planes form an approximate waveguide for signal propagation. Besides the inherited advantages of high-power handling capability and high operating frequency, SIW also shows the benefits of low weight, cost-effective, and high-density integration, etc. However, additional loss is introduced by the dielectric loss inside the waveguide and the leakage loss caused by the separated vias. In addition, SIW exhibits a cutoff frequency
of low values due to its waveguide structure, leading to the limited frequency application range. Acoustic-wave resonators, such as those based on surface-acoustic-wave and bulk-acoustic-wave phenomena, have been identified as a popular RF technology for the mobile front-end stages of wireless communication transceivers due to their stringent requirements for size compactness, low insertion loss, high selectivity, and good linearity. They are usually used in conventional ladder, lattice, or self-cascade configurations to provide high quality factor filtering responses but with limited bandwidth. Hybrid integration schemes in which acoustic-waves resonators are combined with lumped elements are then widely employed to provide better filtering performance. However, the acoustic-wave resonators are typically fabricated by complicated lithography technologies, and the integration with lumped elements imposes more complexity on the fabrication process, resulting in a low yield rate and high cost. Last but not the least, microstrip lines, far lighter, smaller, and cheaper than traditional waveguides, have been studied for several decades. The microstrip line is still by far the most commonly used transmission line structure in RF and microwave designs. The main advantage of microstrip line is the ability to use just two-layer board, with all components mounted on one side, which simplifies both the fabrication and assembly process. While providing the lowest cost RF circuit board solution, microstrip line suffers from radiation issues, crosstalk in high-density routing, signal distortion at high frequency, etc.

The synthesis of multiband filters, whatever the implementation technology, can be done through numerous methods. The simplest method for multiband filter design is to cascade bandpass and bandstop filters [14]. Although it is intuitive and straightforward for implementing a dual-band filter, the cascade connection inevitably leads to greater circuit
size and larger insertion loss. Another popular method is to optimize the coupling matrix for desired multiband filtering response with initial values and optimization algorithms [15]. It may be inefficient under certain circumstances because the convergence is not always guaranteed, and the final topology is difficult to control. Thus, the most popular way to achieve multiband filtering response is utilizing the fundamental frequency and the controllable first harmonic of stepped impedance resonators [16], or stub-loaded multimodal resonators [17]. These resonators can have ring or stub-loaded shapes or consist of quarter- or half-wavelength lines. Each resonator has one or more resonance frequencies, making it possible to form multiband responses by coupling these resonators together. Based on this, a filtering structure consisting of two half-wavelength resonators and two open-stub loaded resonators is first proposed in this dissertation to generate two third-order passbands with high selectivity. This structure is then used to design multiband and multifunctional RF filters to meet the ever-increasing requirements of modern communication systems, such as good signal selectivity and enhanced functions. To demonstrate the efficacy of the proposed filtering structure, it is applied to design a PIN-diodes switch-controlled reconfigurable dual-band bandpass filter (BPF) with four-state filtering responses: both passbands ON, both passbands OFF, high-frequency passband ON, and low-frequency passband ON. In addition, a dual-functional filtering balun is designed by placing two identical filtering structures symmetrically and adding double-sided parallel-strip lines (DSPSLs). The superior performance of these devices suggests that the proposed filtering structure provides a good solution for the design of highly selective filters to make the best use of the limited spectrum resources and reject the interfering signals in wireless systems.
Compared to multiband RF components, tunable RF designs provide a more flexible and easy way to improve multiband wireless communication systems. As given in Figure 1.6, great efforts have been spent to develop tunable RF components with technologies such as mechanical tuning elements [18], semiconductor varactors [19], microelectromechanical systems (MEMS) [20], ferroelectric [21], and ferromagnetic [22] films. Mechanical tuning technique usually tunes the component’s operating frequency by changing a physical dimension. High-Q and high power-handling capabilities can be easily
achieved, but with a bulky size that may lead to difficulty in integration and very low tuning speed. Another more popular tuning method is the application of semiconductor varactors such as PIN diodes and varactor diodes, offering controllable junction capacitance with the applied reverse voltage. The capacitance of a varactor can be controlled by applying a reverse voltage across its P-N junction, which helps to shift the operating frequency of a specific circuit design. Compared to mechanical tuning, semiconductor varactors are superior in integration, cost, and tuning speed (~ns), but suffer from low linearity, power handling capability, and the complexity caused by the introduction of an additional bias circuit. With the rapid development of manufacturing techniques, RF MEMS has been widely applied in the development of numerous components. MEMS devices act as switches or variable capacitors under different bias conditions. Advantages of MEMS, such as extremely high linearity, near-zero power consumption, and low loss, etc., contribute to the wide utilization in designing tunable RF components. However, today’s RF MEMS devices are predominantly based on electrostatic actuation which results in several undesirable effects including high probability of stiction and low power handling capability. In addition, the digital tunability and reliability issues also limit its applications. Ferroelectric or ferromagnetic materials have also been studied to accomplish tunable components with the main advantages of continuous tunability and low response time (~ns). Ferroelectric materials, such as Barium Strontium Titanate (BST) and Lead Zirconate Titanate (PZT), provide controllable and high capacitance density utilizing their high dielectric constant which can be varied with applied voltage. BST has been commonly used to control the frequency and/or phase response of various devices. While offering high capacitance ratios, such BST varactors suffer from low power handling capability and
inherently large non-linearity. Ferromagnetic materials, including cobalt, permalloy (Ni$_{80}$Fe$_{20}$, Py), etc., have been explored in developing RF components with inductive tuning capability. However, they generally require complex external magnetic bias fields and their applications have been limited by the low ferromagnetic resonance (FMR) frequency (<1 GHz).

With high and tunable permeability, ferromagnetic materials are promising in designing tunable RF devices after overcoming the main issue of external biasing magnetic field and low FMR frequency. Methods of increasing FMR frequency have been studied with patterning ferromagnetic thin film [23], increasing film thickness [24], or using ferromagnetic nanowires [25]. As reported in [26], patterning ferromagnetic thin film with a high aspect ratio is an efficient and flexible strategy for high-frequency RF applications with reduced losses. In addition, the permeability of ferromagnetic thin film patterns can be tuned by the magnetic field generated by biasing DC current. Therefore, a lot of tunable devices have been developed by utilizing the patterned ferromagnetic thin films, including tunable inductors [27], tunable bandpass filters [28], tunable phase shifters [29], etc. However, the frequency tunability of previous devices requires a specific design process for the individual component, coupling with a customized fabrication process. A more cost-effective, flexible way to design arbitrary tunable RF components is using the engineered substrate with tunable permeability [30]. The engineered substrate is formed by selectively patterning Py thin films on a normal RF substrate to increase the substrate permeability. In addition, the permeability can be tuned by biasing DC current on the gold lines beneath the Py patterns. By designing arbitrary RF passives on the engineered substrate, good miniaturization, tunable operating frequency, and improved performance are achieved.
This ferromagnetic thin film enabled engineered substrate provides an extra complementary approach of design tunable RF components from the perspective of smart substrate material. Nevertheless, current studies only show the preliminary permeability of one layer of the Py thin film. Although higher permeability and wider tuning range could be achieved with multiple or thicker layers of Py thin films in the engineered substrate, it is time-consuming, costly, and inefficient to fabricate and measure a lot of engineered substrates to find out the specific configuration for a desired permeability. Thus, an accurate EM model of the engineered substrate is highly required to determine the configuration of the engineered substrate quickly and accurately for a specific property of the substrate (e.g., permeability, loss). In this dissertation, the characteristics of the engineered substrate enabled with ferromagnetic thin films are fully investigated and an accurate model is developed to describe the performance or predict the configuration of the engineered substrate.

1.1.2 Signal Integrity in High-Density and High-Data-Rate Systems

A high-density system is achieved by integrating the aforementioned tunable, multiband, and multifunctional RF components, however, a critical issue that comes with it is poor signal integrity. Driven by the ever-evolving requirements of high density and fast signal transmission, there has been a progressively expanded tendency of increasing the density of signal traces, such as microstrip lines, that are usually used for the signal transformation between different components. Within a limited space, signal traces at a high data transfer rate along with other closely placed active traces are vulnerably subject to electric and magnetic coupling, which is known as crosstalk. With the continuous increase of signal switching frequency and bandwidth requirements, the crosstalk,
especially the far-end crosstalk (FEXT), has already become one of the dominant limiting factors for achieving higher data transfer rate.

The most straightforward way to reduce the FEXT is to widen the spacing between two traces. In [31], the trace separation between two adjacent transmission lines is numerically investigated and a spacing rule is developed to describe the function between the trace separation and the trace width. This is not attractive due to the increased routing area and cost. The FEXT can also be mitigated by replacing microstrip with stripline or

Figure 1.7 Technologies for crosstalk reduction: (a) covering dielectric film [32], (b) inserting guard trace [33], (c) introducing groove (groove thickness is described by $d$) [36], (d) using decoupling capacitance [37], (e) adding tabs [38].
similarly adding a layer of dielectric film to the microstrip line [32], as shown in Figure 1.7 (a). However, this is not doable in many practical designs and introduces drawbacks in terms of cost, complexity, and integration. As depicted in Figure 1.7 (b), adding a conductor guard trace between two microstrips is another existing solution, and the guard trace can be a straight line with one or two ends shorted [33], via-stitch guard [34], and serpentine guard [35], etc. Nevertheless, the guard trace requires shorting with ground vias at the endpoints of the microstrips, which can cause issues in practical design. Other technologies, such as introducing groove (Figure 1.7 (c)) [36] or using decoupling capacitors (Figure 1.7 (d)) [37] between the adjacent lines, are employed to reduce the coupling while suffering from complicated and expensive implementation. To overcome this, short trapezoidal-shaped tabs are added to the edges of the wires, as illustrated in Figure 1.7 (e) [38]. These tabs effectively increase the mutual capacitance between the lines without significantly increasing the mutual inductance, and accordingly mitigate the FEXT. Since the added tabs can be fabricated easily together with the original traces, placing little burden on the manufacturing process, this tabbed routing method has been widely used in practical applications by Intel Corporation. Although the mutual capacitance can be improved by optimizing the design variables associated with the tabbed lines, there is still a limitation due to the surface tab routing itself and manufacturing capability. Technologies to further reduce the FEXT are highly desired for modern integrated systems with higher and higher data rates.

In this dissertation, the FEXT of coupled lines and the aforementioned tabbed routing structure are fully analyzed and closed-form formulas are derived to describe the performance. Based on this, thin film techniques are applied to the tabbed routing structure
to further mitigate the FEXT. Since the FEXT occurs when the capacitive coupling between adjacent signal lines is smaller than the inductive coupling, methods can be developed to either increase the capacitive coupling or decrease the inductive coupling. Therefore, three methodologies are employed on the tabbed routing structures to further eliminate the FEXT, e.g., increasing the thickness of the tabs to increase the capacitive coupling, adding ferroelectric thin films with high permittivity between the tabs to increase the capacitive coupling, and depositing ferromagnetic thin films with high permeability on the metal lines to decrease the inductive coupling. With these techniques, signal integrity of high-speed integrated circuits and communication systems is greatly improved.

1.2 Chapter Outline

Following the motivation and literature review demonstrated in the previous sections, the primary objective of this research is to develop miniaturized, multifunctional, and reconfigurable RF technologies with the consideration of signal integrity to accelerate the development of modern wireless communication systems. Accordingly, this dissertation is organized as follows:

Chapter 2 first proposes a filtering structure consisting of two half-wavelength resonators and two open-stub loaded resonators, which generates two third-order passbands. Multiple transmission zeros are introduced by the newly developed coupling scheme, resulting in extremely sharp roll-off desirable for highly selective filters. This filtering structure is applied to design a PIN-diodes switch-controlled reconfigurable dual-band bandpass filter (BPF) with four-state filtering responses: both passbands ON, both passbands OFF, high-frequency passband ON, and low-frequency passband ON. Stepped-
impedance open stubs and one-end-grounded coupled lines are studied and employed to suppress unwanted responses. In addition, two identical filtering structures are placed symmetrically to design a dual-band balun BPF. Double-sided parallel-strip lines (DSPSLs) are added to the input port of the filter, and their inherent out-of-phase feature enables the balun filter to convert unbalanced signal to balanced signal at desired frequencies. The excellent performances of the implemented filter and filtering balun successfully demonstrate the advantages and design efficacy of the proposed filtering structure for different applications.

Chapter 3 introduces a permalloy-enabled engineered substrate with increased and tunable permeability. The electrical tuning mechanism, strategies for improving FMR frequency and magnetic loss are studied and discussed. Based on current experiment results, the engineered substrate with one-layer Py thin films only achieves a limited increase in substrate permeability. More or thicker ferromagnetic thin film layers, not limited to Py thin films, are required to further improve the effective permeability of the entire substrate. Thus, an accurate model is developed to quickly optimize and determine the configuration of the engineered substrate for a desired permeability. The model is generated by thoroughly exploring the performance of the engineered substrate on the effects of number of film layers, film thickness, film vertical locations, planar filling density, and pattern dimension. Conformal mapping theory is employed to determine the effective area of ferromagnetic patterns for different test structures on the engineered substrate. Moreover, the relationship between the effective permeability and the inductance of microstrip lines is extracted for quick modeling. Eventually, the proposed engineered substrate model is derived and verified with numerous test cases. To demonstrate the
design efficacy, a miniaturized and tunable frequency selective surface (FSS) and a miniaturized and performance enhanced patch antenna are implemented on the engineered substrate optimized by the developed engineered substrate model.

Chapter 4 discusses improving signal integrity of coupled microstrip lines through thin film techniques. When two microstrip traces get closer to each other, the increased far-end crosstalk (FEXT) degrades severely the received signal on the same trace. Such FEXT is mitigated by introducing tab routing into microstrip lines while the introduced tabs bring limited mutual capacitance, resulting in the finite reduction of FEXT. To further eliminate the FEXT, the FEXTs of coupled lines and tabbed routing structure are first analyzed by deriving closed-form formulas of capacitance and inductance. Based on the theoretical analysis, three improved methodologies are proposed: increasing the tab thickness and keeping the main lines unchanged, introducing high permittivity ferroelectric materials to the selected area (gap) between the tabs, and covering all the metal lines with high permeability ferromagnetic film. The measured results of the samples enabled by these three methodologies all show great improvement in the signal integrity with FEXT close to zero. By applying these useful methodologies to high-density routing traces, the overall signal integrity of highly integrated wireless communication systems is improved.

Chapter 5 firstly gives a summary of contributions to the dissertation and then presents future work to be performed.
CHAPTER 2
A DUAL-BAND FILTERING STRUCTURE FOR HIGHLY SELECTIVE RECONFIGURABLE BANDPASS FILTER AND FILTERING BALUN

2.1 Introduction

The vigorous development of communication technologies has laid a solid foundation for various applications such as autonomous driving, smart home appliances, and industrial IoTs. However, current and future wireless communication systems are demanded for technical improvement to support multiple wireless communications bands and protocols. Multiband cognitive radio has been widely considered as one of the prominent solutions to tackle the spectrum scarcity with significantly enhanced throughput and better channel maintenance of communication networks [39]. Multiband and multiple function components such as bandpass filters (BPFs) are of great importance in improving system spectral efficiency, system space-consuming, and multi-function capability. Yet, the rapid development of system integration requires multiple filters to work simultaneously within narrow proximity bands, resulting in great demand of filters with high selectivity to make best use of the limited spectrum resources and to reject the interfering signals [40].
To meet the ever-increasing requirements of multi-band and multi-function communication systems, BPFs are required to provide good signal selectivity and enhanced functions simultaneously. Figure 2.1 (a) gives a typical configuration of a front-end module. Simplified module enabled with (b) dual-band switchable filter and (c) dual-band balun filter.

Figure 2.1 (a) Typical configuration of a front-end module. Simplified module enabled with (b) dual-band switchable filter and (c) dual-band balun filter.

To meet the ever-increasing requirements of multi-band and multi-function communication systems, BPFs are required to provide good signal selectivity and enhanced functions simultaneously. Figure 2.1 (a) gives a typical configuration of a front-end...
module, where balun, switches, power amplifier (PA), low-noise amplifier (LNA) and BPFs are employed to enable multi-channel communications. If the BPF is designed to achieve reconfigurable dual passbands, which provides adaptive filtering response to accommodate changes in the available spectrum, the entire module will be significantly simplified as shown in Figure 2.1 (b) [41][42]. Technologies, such as combing stub-loaded resonators with transition structures [43], using dual-frequency resonators [44] or connected-coupling lines [45] and employing stepped-impedance open stubs [46], etc., have been developed to achieve switchable dual-band filtering responses. However, maintaining sharp roll-off, achieving low insertion loss at ON-state, and providing deep signal suppression level at OFF-state, are key technical challenges in designing switchable dual-band BPFs. Similarly, as depicted in Figure 2.1 (c), the size and cost of the module are reduced by integrating dual-band BPFs with balun function. Balun filters convert unbalanced signals to balanced signals and provide filtering responses simultaneously. Although filtering balun structures have been widely achieved with microstrip branch lines [47], substrate integrated waveguides [48], stub loaded resonators [49], and dielectric resonators [50], the integrated technologies for balun function inevitably deteriorate the performance of filters, including poor selectivity and extra insertion loss [51]. Thus, lower loss, smaller size, and higher selectivity are consistent pursuits in the design of multi-functional dual-band BPFs. Nevertheless, high selectivity of dual-band BPFs is hard to achieve with regular filtering structures due to the limited transmission zeros and coupling restrictions. Designing dual-band BPFs with sharp roll-off and good out-of-band rejection remains a challenging task, especially for BPFs with frequency reconfigurability or balun functions.
In this section, a novel filtering structure enabled with half-wavelength resonators and open-stub loaded resonators is presented to provide high-order passbands. By properly designing the coupling scheme, multiple transmission zeros are introduced to achieve extremely high selectivity at passbands. Based on the proposed structure, a PIN-diode-controlled reconfigurable dual-band BPF and a dual-band filtering balun are designed and implemented to demonstrate the design efficacy. The working principles of the basic filtering structure are fully investigated with a thorough theoretical analysis. Measured results of the PIN-diode controlled reconfigurable BPF are presented. Furthermore, double-sided parallel-strip lines (DSPSLs) are integrated into the proposed filtering structure to develop a filtering balun with balanced filtering response. The measured results of the balun filter fully demonstrate the advantages of designing highly selective and multifunctional BPFs with the proposed filtering structure.

2.2 Basic Dual-Band Filtering Structure

Figure 2.2 (a) Simplified open-stub loaded resonator and its equivalent circuit under (b) odd-mode, and (c) even-mode.

Open-stub loaded resonator is widely utilized in the design of miniaturized filter due to its inherent two resonant modes [52]. Two transmission poles are achieved with one resonator and their frequencies are highly related to the line length, leading to the
controllable performance. To obtain better miniaturization and flexible design, meander line technology is employed to optimize such kind of resonator. As shown in Figure 2.2 (a), a simplified open-stub loaded resonator consists of a half-wavelength resonator and an open-stub resonator. The lengths and the characteristic impedances of these two resonators are represented by \((L, Z)\) and \((L’, Z’)\), respectively. The two resonant modes of such open-stub resonator are analyzed with the odd-even method. In odd-mode, the resonator is performed as a transmission line \((L/2)\) with one-end-grounded and its equivalent circuit is shown in Figure 2.2 (b). When the resonance condition of \(Y_{in, odd} = 0\) is met, the first odd-mode resonance occurs, and its resonant frequency is:

\[
f_{odd} = \frac{c}{2L\sqrt{\varepsilon_{eff}}} \tag{2.1}
\]

where \(c\) is the speed of light in the free space and \(\varepsilon_{eff}\) is the effective permittivity of the substrate. Obviously, by controlling the \(L\) value, the odd-mode resonant frequency can be shifted. On the other hand, in even mode, the resonator’s is considered as a half-wavelength resonator with an equivalent circuit illustrated in Figure 2.2 (c). When the impedance of the open-stub \(Z’’\), whose width is half of the open-stub in Figure 2.2 (a), is equal to the impedance of the half-wavelength resonator, i.e., \(Z = Z’’\), the second even-mode resonant frequency is derived as:

\[
f_{even} = \frac{c}{(L+2L)\sqrt{\varepsilon_{eff}}} \tag{2.2}
\]

This means that both \(L\) and \(L’\) affect the even-mode resonant frequency. By changing the length of \(L\) and \(L’\) separately, independent control of the two transmission poles is realized, which provides a high degree of design freedom for reconfigurable filters.
With the integration of the open-stub loaded resonators and half-wavelength resonators, a miniaturized filtering structure with a third-order dual-band filtering response is achieved. As shown in Figure 2.3 (a), two open-stub loaded resonators are placed symmetrically with two half-wavelength resonators and are configured in meander line shape for reduced area. The coupling scheme of this filtering structure is illustrated in Figure 2.3 (b), which shows that multiple transmission poles and zeros are introduced. The odd-modes and even-modes of the two open-stub loaded resonators are denoted by $1^o$ and $1^e$, $2^o$ and $2^e$, respectively. The two half-wavelength resonators 3 and 4 provide signal propagation path between the two open-stub resonators. In odd-mode, resonators $1^o$, $2^o$, and 3 generated three transmission poles to form a low-frequency passband. For even-mode excitation, a high-frequency passband is generated with resonators $1^e$, $2^e$, and 4. With the two separate transmission paths of the passbands, the passbands’ properties, e.g., frequency location, insertion loss, and bandwidth, can be controlled independently. Besides the transmission poles, the source-load coupling between the input and output and the cross-
coupling between the resonators within the filtering structure also generate multiple transmission zeros. By carefully locating the transmission poles and zeros, two third-order passbands with sharp roll-off and good out-of-band suppression are eventually realized.

To further characterize the working mechanism of the proposed filtering structure, coupling matrixes \([M]\) are developed to describe the coupling relationships among resonators. Due to the different roles of resonators, the coupling coefficients \(M_{ij}\) in the matrix are different under odd-mode and even-mode. Based on the above analysis, the coupling matrixes at odd-mode and even-mode are derived as:

\[
[M^o] = \begin{bmatrix}
0 & M_{S1} & 0 & 0 & M_{SL} \\
M_{S1} & M_{11} & 0 & M_{1o3} & 0 \\
0 & 0 & M_{22} & M_{2o3} & M_{2oL} \\
0 & M_{1o3} & M_{2o3} & M_{33} & 0 \\
M_{SL} & 0 & M_{2oL} & 0 & 0
\end{bmatrix}
\]  

(2.3)

\[
[M^e] = \begin{bmatrix}
0 & M_{S1} & 0 & 0 & M_{SL} \\
M_{S1} & M_{11} & 0 & M_{1e4} & 0 \\
0 & 0 & M_{22} & M_{2e4} & M_{2eL} \\
0 & M_{1e4} & M_{2e4} & M_{44} & 0 \\
M_{SL} & 0 & M_{2eL} & 0 & 0
\end{bmatrix}
\]  

(2.4)

The values of the coupling coefficients \(M_{ij}\) are determined by the specific dimensions of the structure. Once the structure is defined, the corresponding coupling matrix can be obtained by fitting the initial guess for the optimization [53]. The loop currents grouped in a 5×5 identity matrix \([I]\) are given as:

\[
[-jR + \omega'[W] + [M^{o,e}]][I] = [A][I] = -j[e]
\]  

(2.5)

where \(j^2 = -1\), \([R]\) is a 5×5 matrix whose only nonzero entries are \(R_{11} = R_{55} = 1\), \([W]\) is similar to \([I]\), except that \(W_{11} = W_{55} = 0\). The excitation vector is \([e] = [1, 0, 0, 0, 0]\. The
low-pass prototype frequency $\omega'$ is related to the actual frequency $\omega$ by

$$\omega' = \omega_0/\Delta \omega \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right),$$

where $\omega_0$ is the center frequency of the filtering circuit and $\Delta \omega$ is its bandwidth. Eventually, the transmission and reflection coefficients of the proposed filtering structure are obtained by (load and source resistors = 1):

$$S_{21} = -2j[A^{-1}]_{5,1} \quad (2.6)$$

$$S_{11} = 1 + 2j[A^{-1}]_{1,1} \quad (2.7)$$

The determined model can be applied to optimize the design of filter minimizing time-consuming full-wave simulation.

2.3 Switchable Dual-band Bandpass Filter

Based on the proposed filtering structure, a dual-band BPF is designed by Ansys HFSS with four independently controlled passband states: both passbands ON, both passbands OFF, high-frequency passband ON, and low-frequency passband ON. As
shown in Figure 2.4, the proposed filtering structure in section II is utilized to generate two third-order passbands with high selectivity. Four PIN diodes are employed to control the connections between signal lines and the ground (GND). The diodes $D_{L1}$ and $D_{L2}$ are located at the odd-mode transmission path to control the operating state of the low-frequency passband. Similarly, the diodes $D_{H1}$ and $D_{H2}$ are arranged at the even-mode transmission path to determine the working state of the high-frequency passband. As shown in Table 2.1, when a group of diodes are turned on, the connected signal suppression mechanism will be activated, resulting in the OFF-state of the corresponding passband.

To achieve better OFF-state rejection, two signal suppression techniques are employed in the switchable filter. As shown in Figure 2.5 (a), when one end of a coupled line is grounded, the input signal from port 1 will not pass through the coupled line to the output port 2. This technique is applied to the proposed filter by grounding one end of the signal lines with diodes $D_{L2}$ and $D_{H2}$. Another signal suppression technique of shunting a

<table>
<thead>
<tr>
<th>Activated switches</th>
<th>Filtering states</th>
</tr>
</thead>
<tbody>
<tr>
<td>None</td>
<td>Both bands ON</td>
</tr>
<tr>
<td>$D_{L1}$, $D_{L2}$, $D_{H1}$, and $D_{H2}$</td>
<td>Both bands OFF</td>
</tr>
<tr>
<td>$D_{H1}$ and $D_{H2}$</td>
<td>Low-frequency band ON</td>
</tr>
<tr>
<td>$D_{L1}$ and $D_{L2}$</td>
<td>High-frequency band ON</td>
</tr>
</tbody>
</table>

![Figure 2.5](image)

Figure 2.5 Signal suppression technologies utilized in the filter: (a) one-end-grounded coupled line, (b) stepped-impedance open stub.
stepped-impedance open stub to a transmission path is also applied in this filter, as illustrated in Figure 2.5 (b). The introduced stub generates a notch frequency response, thereby enhances the signal suppression. The operating frequency of the notch response is mainly determined by the total length of the open stub while the bandwidth is affected by the ratios of $Zs_1$ and $Zs_2$. Moreover, when the open stub is grounded, the influence of the notch response is eliminated. Therefore, two stepped-impedance open stubs are utilized in the proposed filter for each passband, and the working states of the filter are controlled by

Table 2.2: Parameter values used in the switchable BPF

<table>
<thead>
<tr>
<th>Param.</th>
<th>$L_1$</th>
<th>$L_2$</th>
<th>$L_3$</th>
<th>$L_4$</th>
<th>$L_5$</th>
<th>$L_6$</th>
<th>$L_7$</th>
<th>$L_8$</th>
<th>$L_9$</th>
<th>$L_{10}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>(mm)</td>
<td>10.2</td>
<td>12</td>
<td>5</td>
<td>16</td>
<td>9</td>
<td>5.2</td>
<td>11</td>
<td>8.8</td>
<td>19.9</td>
<td>6.6</td>
</tr>
<tr>
<td>$L_{11}$</td>
<td>11.1</td>
<td>5.5</td>
<td>5.5</td>
<td>3.1</td>
<td>3.1</td>
<td>2.65</td>
<td>1</td>
<td>0.3</td>
<td>0.8</td>
<td>1.6</td>
</tr>
<tr>
<td>$W_7$</td>
<td>$W_8$</td>
<td>$W_9$</td>
<td>$W_{10}$</td>
<td>$W_{11}$</td>
<td>$g_1$</td>
<td>$g_2$</td>
<td>$d_1$</td>
<td>$d_2$</td>
<td>$d_3$</td>
<td>$d_4$</td>
</tr>
<tr>
<td>1</td>
<td>0.3</td>
<td>4</td>
<td>0.3</td>
<td>3.2</td>
<td>1</td>
<td>3.8</td>
<td>0.4</td>
<td>3.9</td>
<td>1</td>
<td>1</td>
</tr>
</tbody>
</table>
the diodes $D_{L1}$ and $D_{H1}$. Eventually, the diodes $D_{L1}$ and $D_{L2}$, $D_{H1}$ and $D_{H2}$ are used to switch the states of low-frequency band and high-frequency band, respectively.

(a)

(b)
The optimized switchable dual-band BPF is fabricated on a F4B ($\varepsilon_r = 2.65$, $\tan\delta =$)

Figure 2.7 Simulated and measured $S$-parameters of the designed switchable BPF: (a) both passbands ON, (b) both passbands OFF, (c) low-frequency passband ON, (d) high-frequency passband ON.

The optimized switchable dual-band BPF is fabricated on a F4B ($\varepsilon_r = 2.65$, $\tan\delta =$
0.0035) substrate with a thickness of 1mm. The values of the parameters in Figure 2.4 are shown in Table 2.2. Figure 2.6 shows optical image of a fabricated filter with SMP1322-079 PIN diodes from Skyworks selected as control switches. The insert in Figure 2.6 also illustrates the biasing circuit for each diode, which is formed with an RF choke inductor $L = 180$ nH, a bypass capacitor $C = 33$ pF, and a resistor $R = 100 \, \Omega$, respectively.

The fabricated sample is measured by Rohde & Schwarz ZVA 67, and the results comparisons with four-state filtering responses are depicted in Figure 2.7 (a) - (d). When all the diodes are turned off, an original filter with two third-order passbands is presented in Figure 2.7 (a). The center frequencies of the two passbands are measured at 2.85 GHz and 4.86 GHz, respectively. The first passband has a -3 dB bandwidth of 174 MHz while the -3 dB bandwidth of the second passband is 291 MHz. The minimum insertion losses of the two passbands are 2.3 dB and 2.0 dB, which are slightly higher than the simulated results. The additional loss is contributed by the loss from PIN diodes and SMA connectors. Multiple transmission zeros contribute to the high selectivity and good out-of-band rejection. The minimum roll-off from -3 dB to -15 dB is measured as $1.70 \times 10^3 \, \text{dB/decade}$ and the signal suppression level out of the passbands is better than -20 dB. Instead, when all the diodes are turned on, both passbands will become stopbands due to the activated signal suppression technologies. The filtering response of all frequency blocking is shown in Figure 2.7 (b), where signal rejections are better than -18.7 dB. Filtering response of the switchable BPF with only one passband at low frequencies is shown in Figure 2.7 (c). The passband at high frequencies is disabled by turning on diodes $D_{H1}$ and $D_{H2}$, where the maximum rejection level is found as -19.7 dB. Similarly, Figure 2.7 (d) shows the filtering response with only one passband at high frequencies. Another passband at low frequencies
is blocked by turning on diodes D_{L1} and D_{L2} while the minimum rejection level is measured as -18.2 dB.

Table 2.3 compares the performance of the designed switchable BPF with other similar filters. Most transmission zeros are achieved in the proposed filter, which contributes to the highest selectivity and very good out-of-band suppression. High-order passbands with good miniaturization are realized with multiple meander line resonators in the design. The bandwidth of each band can be increased by slightly separating the transmission poles, which are realizable with different electrical length of the corresponding resonators. In addition, the filter dimension can be further reduced through layout optimization, such as multilayer design.

<table>
<thead>
<tr>
<th>Refs</th>
<th>$f_1/f_2$ (GHz)</th>
<th>IL (dB)</th>
<th>TZs</th>
<th>Orders</th>
<th>Selectivity (dB/decade)</th>
<th>Size ($\lambda_g^2$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[42]</td>
<td>1.5/2.0</td>
<td>3.1/3.1</td>
<td>3</td>
<td>3/3</td>
<td>4.2 × 10^2</td>
<td>0.40</td>
</tr>
<tr>
<td>[43]</td>
<td>2.51/3.52</td>
<td>1.2/1.3</td>
<td>5</td>
<td>1/1</td>
<td>6.7 × 10^2</td>
<td>0.38</td>
</tr>
<tr>
<td>[44]</td>
<td>0.9/2.16</td>
<td>2.6/3.2</td>
<td>1</td>
<td>2/2</td>
<td>2.6 × 10^2</td>
<td>0.17</td>
</tr>
<tr>
<td>[45]</td>
<td>1.5/2.5</td>
<td>3.0/2.6</td>
<td>2</td>
<td>2/2</td>
<td>4.0 × 10^2</td>
<td>0.27</td>
</tr>
<tr>
<td>[46]</td>
<td>0.9/2.35</td>
<td>1.4/4.9</td>
<td>0</td>
<td>2/3</td>
<td>5.1 × 10^2</td>
<td>0.07</td>
</tr>
<tr>
<td><strong>Proposed</strong></td>
<td><strong>2.85/4.86</strong></td>
<td><strong>2.3/2.0</strong></td>
<td><strong>6</strong></td>
<td><strong>3/3</strong></td>
<td><strong>17.0 × 10^2</strong></td>
<td><strong>0.16</strong></td>
</tr>
</tbody>
</table>

IL: Insertion loss; TZs: Transmission zeros; $\lambda_g$: Guided wavelength at the first frequency band.
2.4 Dual-Band Bandpass Filtering Balun

Compact, efficient, and low-cost signal processing is highly required in modern communication systems. Baluns are commonly used in these systems to convert the

![Diagram of Dual-Band Bandpass Filtering Balun](image)

Figure 2.8 (a) 3D view of the dual-band balun BPF. (b) Schematic layout of the top layer.

33
unbalanced signal to the balanced signal between multiple devices, such as filters, amplifiers, and antennas, etc. The integration of balun and filter can significantly reduce the cost and size of functional blocks, which makes the balun BPF more attractive in recent years. In this section, a dual-band balun BPF based on the presented filtering structure is designed for high selectivity. The symmetrical and planar configuration of the filtering structure makes it suitable for balun design by integrating them back-to-back. As shown in Figure 2.8, the filtering balun consists of five layers: top resonator, top substrate, middle conductor plane, bottom substrate, and bottom resonator. Two filtering structures are applied on the top and bottom layers to provide dual-band filtering responses. At the input and output of the filtering structure, two extra metal lines are added for signal diversion. The top layout and the bottom layout are centrosymmetric along the centerline. The conductor plane behaves as a virtual ground for microstrip lines on both sides. It is noted

![Figure 2.9 Photograph of the fabricated balun BPF: (a) front view, (b) back view.](image)
that the ground layer is partially defected, where a double-sided parallel-strip line (DSPSL) is formed. The inherent out-of-phase feature of DSPSL makes the filter easy to output balanced signal. Since the filtering structures are excited simultaneously at port 1 on both sides, the input energy is equally separated to port 2 and port 3 with a 180° phase difference.

According to [54], when the DSPSL has the same width as the converted microstrip line, the characteristic impedance of DSPSL will be twice that of the converted microstrip lines. This property shows great advantages of DSPSL in high impedance applications and enables the easy conversion between the DSPSL and the back-to-back microstrip lines. Eventually, a dual-band balun BPF is constructed to generate a balanced filtering response. The theory in section II is used to analyze and optimize the filtering balun. Since the designed balun BPF consists of two layers of identical filtering structure, the signal at two output ports is half of the calculated results from equation (3.6), i.e., the calculated transmission coefficients should minus 3 dB for the balun BPF. In addition, the inverse output phases at port 2 and port 3 should be considered during the calculation.

<table>
<thead>
<tr>
<th>Param.</th>
<th>(L_1)</th>
<th>(L_2)</th>
<th>(L_3)</th>
<th>(L_4)</th>
<th>(L_5)</th>
<th>(L_6)</th>
<th>(L_7)</th>
<th>(L_8)</th>
<th>(L_9)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(mm)</td>
<td>7</td>
<td>14</td>
<td>16</td>
<td>5.2</td>
<td>9</td>
<td>11</td>
<td>8.9</td>
<td>19.9</td>
<td>6.4</td>
</tr>
</tbody>
</table>

As shown in Figure 2.9, a well-designed balun BPF is fabricated on F4B \((\varepsilon_r = 2.65,\ \tan\delta = 0.0035)\) substrate with dimensions listed in Table 2.4. The centrosymmetric structure is realized by bonding two identical printed circuit boards (PCBs) with silver glue.
In order to solder SMA conductors to port 2 and port 3, the PCBs are made partially open to expose the intermediated conductor plane without affecting the filter performance.

Figure 2.10 Simulated and measured results of the balun BPF: (a) $S$-parameters, (b) phase and magnitude imbalances of each passband.
The simulated and measured results of the balun BPF are presented in Figure 2.10. Almost the same filtering responses are generated at port 2 and port 3, which fully validate the design efficacy of the proposed balun BPF. The measured center frequencies of the two passbands are 2.89 GHz and 4.81 GHz, respectively. The first passband has a -3 dB bandwidth of 100 MHz while the -3dB bandwidth of the second passband is 180 MHz. Extremely high selectivity of the two passbands is achieved by the proposed coupling scheme and the minimum roll off from -3 dB to -15 dB is $2.71 \times 10^3$ dB/decade. The minimum insertion losses at two passbands are measured to be (1.77 + 3) dB and (1.41 + 3) dB. The additional +3 dB in the insertion losses is because the input power is equally divided into the two output ports. The extra loss compared to the simulated results is mainly caused by the manufacturing variation from silver gluing and SMA connectors losses. In addition, the measured transmission zeros are slightly shifted from the simulated $S$-parameters, which are generated by the alignment errors during the bonding process and the use of thicker intermediate conductor plane. To further show the balanced filtering

<table>
<thead>
<tr>
<th>Refs</th>
<th>$f_1/f_2$ (GHz)</th>
<th>IL (dB)</th>
<th>FBW (%)</th>
<th>Selectivity (dB/decade)</th>
<th>Size ($\lambda_g^2$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[47]</td>
<td>2.85/3.15</td>
<td>1.9/1.7</td>
<td>5.2/5.1</td>
<td>$1.9 \times 10^2$</td>
<td>1.95</td>
</tr>
<tr>
<td>[48]</td>
<td>9.51/15.0</td>
<td>2.8/2.4</td>
<td>2.7/5.2</td>
<td>$19.7 \times 10^2$</td>
<td>0.96</td>
</tr>
<tr>
<td>[49]</td>
<td>2.44/3.50</td>
<td>2.1/2.2</td>
<td>4.9/2.7</td>
<td>$3.6 \times 10^2$</td>
<td>0.09</td>
</tr>
<tr>
<td>[50]</td>
<td>1.52/1.64</td>
<td>0.9/0.9</td>
<td>0.8/0.9</td>
<td>$14.0 \times 10^2$</td>
<td>1.56</td>
</tr>
<tr>
<td>[51]</td>
<td>3.93/4.14</td>
<td>0.5/0.7</td>
<td>1.0/1.1</td>
<td>$22.7 \times 10^3$</td>
<td>0.32</td>
</tr>
<tr>
<td><strong>Proposed</strong></td>
<td><strong>2.89/4.81</strong></td>
<td><strong>1.8/1.4</strong></td>
<td><strong>3.5/3.7</strong></td>
<td><strong>27.1 \times 10^2</strong></td>
<td><strong>0.18</strong></td>
</tr>
</tbody>
</table>

IL: Insertion loss; $\lambda_g$: Guided wavelength at the first frequency band.
response of the proposed balun BPF, Figure 2.10 (b) gives the phase and magnitude imbalances of each passband. The maximum imbalances of phase and magnitude are measured as 0.3 dB and 5 °, respectively.

Table 2.5 summarizes the performance comparison of the balun BPF with other similar filters. Due to the additional transmission zeros introduced by the novel coupling scheme, high selectivity and excellent out-of-band rejection are achieved. Adoption of multiple resonators and proper design make the designed balun BPF has high-order passbands and good miniaturization factor. As described in the design of the switchable BPF, the bandwidth of the two passbands for the presented filtering balun can be further increased by slightly separating the transmission poles.

2.5 Conclusion

This chapter proposes a novel highly selective filtering structure with a carefully designed coupling scheme. The filtering structure consists of two half-wavelength resonators and two open-stub loaded resonators, which generates two third-order passbands. Multiple transmission zeros are introduced by the newly developed coupling scheme, resulting in extremely sharp roll-off desirable for highly selective filters. The proposed structure is applied to design a PIN-diodes switch-controlled reconfigurable dual-band bandpass filter with four-state filtering responses: both passbands ON, both passbands OFF, high-frequency passband ON, and low-frequency passband ON. Stepped-impedance open stubs and one-end-grounded coupled lines are studied and employed in the design to suppress unwanted responses. In addition, two filtering structures are placed symmetrically to design a dual-band balun BPF. Double-sided parallel-strip lines are added in the input
port of the filter, and their inherent out-of-phase feature enables the balun filter to convert unbalanced signal to balanced signal at desired frequencies. The measured results of both fabricated devices agree well with the simulated results. The switchable BPF and balun BPF have two passbands (2.8 GHz and 4.8 GHz) with high selectivity with the minimum roll-offs from -3 dB to -15 dB are $1.70 \times 10^3$ dB/decade and $2.71 \times 10^3$ dB/decade, respectively. The insertion loss of the switchable BPF is 2.3 / 2.0 dB and the minimum signal suppressions level OFF state is measured as -18.2 dB. The insertion loss of the balun BPF is 1.8 / 1.4 dB, and the phase and magnitude imbalances are less than 0.3 dB and 5 °, respectively. The excellent performance of the implemented BPF and filtering balun fully demonstrates the design efficacy of the proposed filtering structure for reconfigurable and multifunctional BPFs with high selectivity.
CHAPTER 3
MODELING OF PERMALLOY ENABLED ENGINEERED SUBSTRATE FOR MINIATRUIZED RF COMPONENTS WITH IMPROVED PERFORMANCE

3.1 Introduction

In modern wireless communication systems, most miniaturized RF passives are achieved on dielectric substrate with high permittivity. Generally, the high permittivity dielectric will cause high capacitive effects between the metal trace and the reference ground. To reduce the coupling effects and keep the desired RF impedance, high inductance from narrow trace is needed to balance the high capacitance, which will bring high loss and coupling noise between traces. Magneto-dielectrics with both high permittivity and permeability ($\varepsilon_{\text{eff}} > 1$, $\mu_{\text{eff}} > 1$) could be properly used to reduce the capacitive effects in the wireless communication system [55]. Based on this, a novel approach of developing magneto-dielectric substrate with patterned permalloy (Py) thin films is reported in [30]. As shown in Figure 3.1 (a), the engineered substrate consists of normal RF substrate, patterned ferromagnetic thin films, and metal bias lines. By selectively patterning the ferromagnetic thin films with a high aspect ratio, the ferromagnetic resonant (FMR) frequency is increased to RF frequencies and the magnetic loss caused by the ferromagnetic thin films is limited. In addition, the permeability of the engineered substrate can be tuned by applying DC current to the metal bias lines.
beneath the ferromagnetic thin films. With tunable permeability and limited substrate loss, the engineered substrate provides a cost-effective and flexible way to design arbitrary miniaturized and tunable RF components [56].

In order to get higher substrate permeability, more layers of ferromagnetic thin films are required, which makes it critical to develop an accurate model to quickly optimize and determine the configuration of the engineered substrate for the desired permeability. Therefore, an accurate model is derived in this section by simulating microstrip line models with different configurations of the engineered substrate. The effects of film thickness, number of film layers, film vertical position, planar filling density, and pattern dimension on the engineered substrate properties are thoroughly explored. Conformal mapping theory is employed to determine the effective area of ferromagnetic patterns under different test structures of various microstrip lines on the engineered substrate. Moreover, the relationship between the effective permeability and the inductance of microstrip lines is extracted for quick modeling. Eventually, the proposed engineered substrate model is
derived and verified with numerous test cases. To demonstrate the design efficacy, the derived model is used to determine the configuration of the engineered substrate for two design examples. A frequency selective surface (FSS) is first designed to achieve good miniaturization and tunable operating frequency presented. In addition, the benefits of the optimized engineered substrate on the antenna performance, such as increasing bandwidth, are discussed.

3.1.1 Ferromagnetic Materials

Magnetic materials have magnetic moments derived from electrons’ rotation and orbital angular moments. Those materials exhibit magnetic polarization when exposed to an applied magnetic field. According to the response to the external magnetic field, magnetic materials are classified into five main groups: diamagnetic, paramagnetic, ferromagnetic, anti-ferromagnetic and ferrimagnetic materials [57]. Magnetic dipoles spontaneously align in the same direction within individual magnetic domain of ferromagnetic materials. When the ferromagnetic material is exposed to a magnetic field, magnetic domains are aligned in the same direction as the bias field and maintain magnetic properties even in the absence of a magnetic field [58]. As illustrated in Figure. 3.2, the

![Figure 3.2](image)

Figure 3.2 (a) Unmagnetized-ferromagnetic, (b) magnetized-ferromagnetic.
A ferromagnetic material is magnetized with its magnetic moments of domains parallel to the external magnetic field, resulting in high and relatively stable permeability. Fe, Ni, Co and their alloys are examples of ferromagnetic materials that are greatly desired in the design of RF components.

3.1.2 Permeability Tuning Mechanism of Ferromagnetic Materials

![Hysteresis loop and electrical tuning method of ferromagnetic materials](image)

Figure 3.3 (a) Hysteresis loop and (b) electrical tuning method of ferromagnetic materials.

When a ferromagnetic material is placed in a magnetic field ($H$), the nonlinear cycle of magnetization and demagnetization is called a hysteresis loop, as depicted in Figure 3.3 (a) [59]. With the applied external magnetic field increasing in the positive direction, the magnetic flux density ($B$) increases nonlinearly and eventually saturates at $B_s$. After saturation, the magnetic induction reduces to $B_r$ when the external magnetic field is removed. On the other hand, when the applied magnetic field is reduced along the negative direction, the magnetic induction is decreased to zero, where the required magnetic field is called coercivity ($H_c$). If the external magnetic field keeps decreasing and becomes a reversed field, the curves go symmetrically about the origin point.
Magnetic permeability ($\mu$) is equal to the magnetic flux density $B$ established within the material by a magnetizing field divided by the magnetic field strength $H$ of the magnetizing field ($\mu = B/H$). The permeability of ferromagnetic materials is thus tunable with an external magnetic field, which is often bulky noisy, requires comparatively high-power consumption for operation, and is difficult to integrate with modern mobile communication systems. To minimize the need of bulk external magnetic bias field, DC current has been utilized to electrically tune the permeability of ferromagnetic thin film [26]. As shown in Figure 3.3 (b), DC current is applied through the patterned metal DC bias lines which tune the magnetization distribution inside ferromagnetic thin films. The maximum magnetic field associated with the applied DC currents is estimated with Ampere’s law:

$$H_{DC} = \frac{I_{DC}}{2W}$$  \hspace{1cm} (3.1)

where $I_{DC}$ is the applied DC current, and $W$ is the width of the metal bias lines.

The relative permeability of ferromagnetic film can then be expressed as:

$$\mu_r = \frac{4\pi M_s}{H_k + H_{DC}} + 1$$  \hspace{1cm} (3.2)

where $M_s$ and $H_k$ represent saturation magnetization and internal induced magnetic field of the ferromagnetic film, respectively. The static magnetic field generated by DC current tilts the magnetization direction in the film away from its easy axis towards the hard axis, which in turn decreases the material’s magnetic moments: saturation magnetization and anisotropy field. The equivalent permeability of the Py thin film is thus tuned accordingly.
3.1.3 FMR Frequency of Ferromagnetic Materials

Ferromagnetic resonance (FMR) is the coupling between an electromagnetic wave and the magnetization of a medium through which it passes, and electromagnetic wave has a peak loss at FMR frequency [60]. Ferromagnetic films have been explored in developing RF components with inductive tuning capability, however, FMR frequency of un-patterned ferromagnetic films falls into the sub-GHz range. The resonant frequency of a ferromagnetic film with parallel applied external field $B$ is given by the Kittel formula [61]:

$$f_{FMR} = \frac{\gamma}{2\pi} \sqrt{(H_{Bias} + H_{Ani} + (N_y - N_z)4\pi M_s)(H_{Bias} + H_{Ani} + (N_x - N_z)4\pi M_s)} \tag{3.3}$$

where $N_x$, $N_y$, and $N_z$ are demagnetization coefficients for the ferromagnetic materials in different directions, $M_s$ is the saturation magnetization, $\gamma$ is the gyromagnetic ratio, $H_{Bias}$ is the applied magnetic field, and $H_{Ani}$ is the self-biased shape anisotropy field.

Patterning ferromagnetic thin film with high aspect ratio is an efficient and flexible strategy for high frequency RF applications with reduced losses [26]. The properly patterned ferromagnetic thin films have built-in high shape anisotropy fields providing a self-biasing field, thus increasing FMR frequency according to Kittel equation (3.3). The patterned ferromagnetic thin films (e.g., Ferrite, Py, CoNbZr, etc.) with higher saturation magnetic field can support applications beyond 20 GHz.

3.1.4 Magnetic Loss of Ferromagnetic Materials

Besides the limitation of FMR frequency, another challenge for the application of ferromagnetic materials is their magnetic loss. Dependent on the properties of the material, there are two different ways by which the energy dissipation take place: one is hysteresis...
loss, and the other is eddy current loss [57]. According to hysteresis loop in Figure 3.3 (a), hysteresis loss is defined as the energy dissipated in the form of heat due to hysteresis. Generally, a narrow area of hysteresis is required for the application of magnetic material in RF components. Eddy current loss refers to the energy dissipated by free electrical charges in a conducting material when the latter is situated in a varying magnetic field. Small or reduced eddy current loss is also highly required in RF applications. The eddy current induced on the thin film surface could generate resistive losses and lower the effective permeability. To suppress eddy current, the thickness of thin film is chosen to be less than the skin depth, and the patterns of thin film are perpendicular to the magnetic direction by decreasing the agglomerated particles. In practical applications, the loss caused by the eddy current of the ferromagnetic material, such as Py, can be eliminated by replacing Py with ferrite, which has the same permeability as Py but zero conductivity.

In general, with high and tunable permeability, ferromagnetic materials are promising in the design of tunable microwave components. Compared to conventional biasing method with external magnetic field, electrically tuning ferromagnetic materials with DC current shows a great advantage in the easy integration with other components on a common fabrication platform.

3.2 Characterization of Permalloy Enabled Engineered Substrate

It has been demonstrated that the magnetic loss of the engineered substrate can be reduced by selectively patterning the ferromagnetic films [62]. In addition, the eddy current loss caused by high-conductivity ferromagnetic materials, such as permalloy, can be eliminated by replacing it with ferrite, which has the same permeability but close to zero
conductivity. Therefore, the main parameter of the engineered substrate that needs to be fully investigated is the effective permeability. As shown in Figure 3.4, the microstrip line model with ferromagnetic thin films is used to characterize the performance of the engineered substrate. Ferromagnetic thin films with the property of $\varepsilon_r = 1$ and $\mu_r = 1000$ are inserted into a normal RF substrate ($\varepsilon_r = 2.2$ and $\mu_r = 1$). The dimensions of the microstrip line are chosen as $w = 1.52\text{mm}$, $h = 0.5\text{mm}$. The length of the microstrip line is chosen as $0.03\lambda_0$, where $\lambda_0$ is the free space wavelength at an operating frequency of 2.5 GHz. A transmission line with such a short electrical length is equivalent to a section of resistance, inductance, capacitance and conductance ($RLGC$) circuit, and the inductance is extracted from simulated $S$-parameters of the microstrip line model in Ansys HFSS [63]. The $S$-parameters are first converted to $ABCD$ parameters and then used to compute the propagation constant $\gamma$ and characteristic impedance by the following equations.

$$\gamma = \frac{sinh^{-1} BC}{l} \quad (3.4)$$

$$Z = \frac{B}{\sqrt{C}} \quad (3.5)$$

where $l$ is the length of the microstrip line. Subsequently, the inductance $L$ is given as:
Figure 3.5 The characteristic impedance of a microstrip line converges at a certain substrate width, which is defined as the effective width where ferromagnetic thin films have significant effects on the extracted inductance. (b) Fitting surface of the effective width ($W_{\text{conv}}$) by least squares method.
Since the extracted inductance of the microstrip line is positively related to the effective permeability of the engineered substrate, it is used to describe the effects of the ferromagnetic thin films.

Before further characterizing the effects of the ferromagnetic thin films on the substrate effective permeability, the effective area of the ferromagnetic thin films should be first determined. The effective area is defined as the area where ferromagnetic thin films have significant effects on the extracted inductance. For the microstrip line model, the effective area is investigated by different effective widths in the insert of Figure 3.5 (a), because the ferromagnetic thin films along the metal line direction obviously have significant effects. When it reaches the effective width, the extracted inductance and the characteristic impedance of the microstrip line converges. For a specific microstrip line model, the effective width is determined by simulating characteristic impedance under different substrate widths and finding the convergence point. Since it is time-consuming to get the impedance under different cases by full-wave simulation, conformal mapping theory is employed in this dissertation to calculate the impedance of microstrip lines with different substrate widths [64]. Figure 3.5 (a) shows the comparison of impedance generated by conformal theory and HFSS simulation, where the same convergence point is found. Hence, effective widths under different dimensions of microstrip lines are quickly calculated by the conformal mapping method, as shown by the blue points in Figure 3.5 (b). Based on this, a fitting surface is generated by the least squares method and the formula of the effective width \((W_{\text{conv}})\) is derived as [65]:

\[
L = Im\{\gamma Z\} 
\]
The error variance of such a fitting formula is 0.025, which can be further reduced by introducing higher-order factors into the fitting formula. With this formula, the effective width of ferromagnetic thin films is directly determined for a given dimension of the microstrip line model.

Ferromagnetic thin films within the effective area will increase the effective permeability of the substrate, resulting in increased inductance. Figure 3.6 shows the characterization process of the engineered substrate enabled by ferromagnetic thin films. The microstrip line with a real engineered substrate is first simulated to extract the increased inductance. On the other hand, an equivalent model with effective permeability of $\mu_{r,\text{eff}}$ is simulated. By increasing the value of $\mu_{r,\text{eff}}$, the equivalent model will achieve the same inductance as the real model, which means the engineered substrate has an effective permeability of $\mu_{r,\text{eff}}$. Such a method can accurately extract the effective permeability of a specific engineered substrate. However, it requires multiple simulations of the equivalent model with different preset values of $\mu_{r,\text{eff}}$ to achieve the same inductance as each specific area.
real model, leading to the inefficient characterization process. It is necessary to develop a closed-form formula to describe the relationship between the extracted inductance and

Figure 3.7 (a) Simulated results of microstrip lines with different dimensions. (b) Comparison of the results generated by HFSS simulation and the fitting formula.
substrate permeability so that the effective permeability can be directly calculated by the simulated inductance of the real model.

A microstrip line with a normal dielectric substrate ($\mu_0 = 1$) has an initial inductance of $L_0$, which is mainly contributed by the metal line. When the normal dielectric substrate is replaced by an engineered substrate with ferromagnetic thin films, the substrate permeability will increase to $\mu_1$ ($\mu_1 = k_{\mu r} \times \mu_0$), resulting in an increase of the extracted inductance to $L_1$ ($L_1 = k_L \times L_0$). Then the relationship between the extracted inductance and substrate permeability is described by a function of $k_L = F (k_{\mu r})$. Figure 3.7 (a) gives the simulated results of hundreds of microstrip line cases under different dimensions ($0.3 \text{ mm} < w < 1.5 \text{ mm}, 0.3 \text{ mm} < h < 1.5 \text{ mm}$). The function of $F (k_{\mu r})$ is then derived from these reference results by considering the dimensions of the microstrip line:

$$k_L = F (k_{\mu r}) = \frac{k_{\mu r}+1}{2} - \frac{k_{\mu r}-1}{2} \times ratio$$  \hspace{1cm} (3.8)

where the factor “ratio” is a fitting formula based on simulated results:

$$ratio = 0.01 \left(\frac{w}{h}\right)^2 - 0.15 b \left(\frac{w}{h}\right) + 0.0014 k_{\mu r}^3 - 0.036 k_{\mu r}^2 + 0.33 k_{\mu r} - 0.4$$  \hspace{1cm} (3.9)

As depicted in Figure 3.7 (b), the derived formula is validated with reference data that was not used in the fitting process. Good agreement is found between the results generated by HFSS simulation and the fitting formula. With this formula, the effective permeability of an engineered substrate with a specific configuration is quickly calculated once the inductance of the microstrip line model is extracted.
After determining the effective area for placing ferromagnetic thin films and the formula for extracting effective permeability of the engineered substrate, the performance of the engineered substrate under different configurations is studied. Figure 3.8 shows the extracted inductance of the microstrip line model under different pattern sizes and orientations while keeping the total area of the ferromagnetic thin films constant. It is noted that the extracted inductance gets close to a stable value when the pattern size is smaller than 0.02% $\lambda_0$, where $\lambda_0$ is the wavelength in free space. In the real application, the ferromagnetic thin films have already been patterned with a size smaller than 0.02% $\lambda_0$ to increase the FMR frequency and reduce the magnetic loss [30]. This means the effects of the ferromagnetic patterns can be ignored and the single layer of ferromagnetic patterns can be considered as an equivalent substrate with increased permeability. Therefore, a multilayer substrate model is developed to calculate the effective permeability of the
engineered substrate [66]. The effective permeability of a multilayer substrate is calculated by:

\[ \mu_{r,eff} = \sum_{n=1}^{N} \frac{\mu_{rn} h_n}{h_t} \times k_{\mu r_1} \times k_{\mu r_n}, n = 1, 2, 3, \ldots \]  

(3.10)

where \( k_{\mu r_1} \ldots k_{\mu r_n} \) are correction factors that describe the additional effects of ferromagnetic thin films in different scenarios. In this dissertation, four dominant correction factors are considered in terms of the vertical position of the ferromagnetic thin film layer (\( g/h \)), film layer thickness (\( t \)), number of film layers, and the planar filling density of the film layer, respectively. The real microstrip line model with the engineered substrate in different scenarios is simulated and the extracted inductance values are used as a reference, as shown by the black lines in Figure 3.9. All of these simulations are conducted within the effective area defined by equation (4). On the other hand, equation (3.10) without correction factors is used to calculate the effective permeability for the equivalent model and the extracted inductance is depicted by the red lines in Figure 3.9. In order to fit the red lines to the black lines, the corresponding fitting functions \( k_{Ln} = f (x) \) are generated. Subsequently, the correction factors \( k_{\mu rn} \) are calculated by equation (3.8) which describes the relationship between \( k_{Ln} \) and \( k_{\mu rn} \). The vertical position of the ferromagnetic thin film layer is first explored by changing the values of parameter \( g \) in Figure 3.9 (a). In the real model, the farther the ferromagnetic layer is, the smaller its effects, while the equivalent model generates a flat line because the vertical position of the film layer is not considered. Thus, the function \( k_{L1} = f (g/h) \) is derived as:

\[ k_{L1} = f \left( \frac{g}{h} \right) = 0.13 \left( \frac{g}{h} \right)^2 - 0.24 \left( \frac{g}{h} \right) + 1.02 \]  

(3.11)
Figure 3.9 (b) simulates the real model and equivalent model with a film thickness of $2t$. 
The film layer is placed at a vertical position where $k_{\mu 1} = 1$ to avoid the influence of the...

Figure 3.9 Derivation of correction factors in different scenarios: (a) vertical position, (b) film thickness, (c) number of film layers, and (d) planar filling density.
vertical position. A good agreement is found between the real model and the equivalent model, since the thickness factor has already been considered in equation (3.10), resulting in the corresponding correction factor $k_{\mu_2}$ equal to 1. Figure 3.9 (c) divides the $2t$ thick film layer into two layers and keeps the top film layer at the same vertical position as Figure 3.9 (b). The results show that the top film layer has a dominant effect on the extracted inductance, which means the correction factor of vertical position for the model with multiple thin film layers is determined by only considering the distance between the first film layer and the top surface. In addition, the model with multiple thin film layers is calculated by substituting the total thickness of these film layers into equation (3.10). All previous cases are simulated by filling the effective area of the microstrip line model with ferromagnetic thin films, which means 100% planar filling density. In Figure 3.9 (d), the planar filling density $\eta$, which is defined as the ratio of the ferromagnetic thin film area and the total effective area ($\eta = A_{\text{ferro}} / A_{\text{eff}}$), is studied by changing the value of parameter $l$. The smaller area of the ferromagnetic thin film, the smaller extracted inductance. Hence, the function of $k_{L3} = f(\eta)$ is fitted as:

$$k_{L3} = f(\eta) = 0.51\eta + 0.51$$  \hspace{1cm} (3.12)

Eventually, the accurate multilayer substrate model is created with equation (3.10) - (3.12). The developed model is used to predict the configuration of an engineered substrate for the desired permeability. In order to demonstrate the design efficacy, the following section will introduce two design examples on the engineered substrate optimized by the proposed model. In addition, the benefits of the engineered substrate for different applications are explored.
A sample of the engineered substrate with a layer of 100 nm thick permalloy thin film has been presented in [62]. The permalloy thin film is patterned as an array of 15 \( \mu \text{m} \times 20 \, \text{\mu m} \) with 10 \( \mu \text{m} \) gaps among them. Gold lines are deposited beneath the permalloy patterns for biasing currents. A simple patch antenna is put on such an engineered substrate to characterize the permeability performance by the shift of operating frequency. As shown in the second column of Table 3.1, the engineered substrate with one layer of permalloy thin film provides tunable effective permeability from 1.140 to 1.102 by applying DC current from 0 mA to 500 mA. By utilizing the proposed multilayer substrate model, the effective permeability of the single 100 nm thick permalloy is calculated as 841.140 without biasing current. Furthermore, increased permeability of 2.398 is achieved by increasing the layers of the permalloy thin film to 10 and the calculated results under different DC currents are also shown in Table 3.1. Based on this, two design examples are presented in the following to demonstrate the design efficacy of the proposed model and further explore the benefits of the ferromagnetic thin film enabled engineered substrate.

### Table 3.1: Effective permeability calculated by the proposed model

<table>
<thead>
<tr>
<th>DC Current</th>
<th>Effective permeability (1 layer)</th>
<th>Permeability of permalloy thin film layer</th>
<th>Estimated effective permeability (10 layers)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0mA</td>
<td>1.140</td>
<td>841.140</td>
<td>2.398</td>
</tr>
<tr>
<td>100mA</td>
<td>1.134</td>
<td>805.134</td>
<td>2.338</td>
</tr>
<tr>
<td>200mA</td>
<td>1.130</td>
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<td>2.298</td>
</tr>
<tr>
<td>300mA</td>
<td>1.123</td>
<td>739.123</td>
<td>2.228</td>
</tr>
<tr>
<td>400mA</td>
<td>1.109</td>
<td>655.109</td>
<td>2.088</td>
</tr>
<tr>
<td>500mA</td>
<td>1.102</td>
<td>613.102</td>
<td>2.018</td>
</tr>
</tbody>
</table>

3.3 Model Verification and Design Examples

A sample of the engineered substrate with a layer of 100 nm thick permalloy thin film has been presented in [62]. The permalloy thin film is patterned as an array of 15 \( \mu \text{m} \times 20 \, \text{\mu m} \) with 10 \( \mu \text{m} \) gaps among them. Gold lines are deposited beneath the permalloy patterns for biasing currents. A simple patch antenna is put on such an engineered substrate to characterize the permeability performance by the shift of operating frequency. As shown in the second column of Table 3.1, the engineered substrate with one layer of permalloy thin film provides tunable effective permeability from 1.140 to 1.102 by applying DC current from 0 mA to 500 mA. By utilizing the proposed multilayer substrate model, the effective permeability of the single 100 nm thick permalloy is calculated as 841.140 without biasing current. Furthermore, increased permeability of 2.398 is achieved by increasing the layers of the permalloy thin film to 10 and the calculated results under different DC currents are also shown in Table 3.1. Based on this, two design examples are presented in the following to demonstrate the design efficacy of the proposed model and further explore the benefits of the ferromagnetic thin film enabled engineered substrate.
3.3.1 Miniaturized and Tunable Frequency Selective Surface (FSS)

Frequency selective surfaces (FSSs) which are usually two-dimensional (2-D) infinite arrays, acting as spatial filters for the incident electromagnetic waves, have been widely investigated over the past decades [67]. As shown in Figure 3.10, a square ring based FSS is implemented on a Roger RT/Duriod 5880 ($\varepsilon_r = 2.2$, $\mu_r = 1.0$, $\tan\delta = 0.0009$) substrate embedded with ten layers of 100 nm thick permalloy thin films. The square loops form a first order bandstop filtering response, and the operating frequency is approximately calculated by:

$$f_0 = \frac{1}{2\pi\sqrt{LC}}$$

(3.13)

where the inductance and capacitance are directly dependent on the permeability ($\mu_r$) and permittivity ($\varepsilon_r$) of the substrate. FSSs are typically miniaturized by increasing the substrate permittivity, resulting in the strong capacitive coupling between adjacent unit cells, which

Figure 3.10 A square ring based FSS on the engineered substrate.
degrades the performance of the FSS. A good solution for this is to increase the substrate

![Diagram](image1.png)

**Figure 3.11** (a) Results comparison of the FSS with and without the engineered substrate. (b) Performance of the miniaturized FSS under different DC biasing conditions.
permeability so that the same miniaturization factor \((n = \sqrt{\mu_r \varepsilon_r})\) is achieved with moderate values of \(\mu_r\) and \(\varepsilon_r\) while the capacitive coupling is reduced.

Therefore, due to the increased effective permeability of the engineered substrate, the same operating frequency is achieved with smaller a unit cell size. By implementing FSS on such an engineered substrate, the unit cell size can be significantly reduced from the original \(0.311\lambda_0 \times 0.311\lambda_0\) to \(0.285\lambda_0 \times 0.285\lambda_0\), where \(\lambda_0\) is the free space wavelength at the operating frequency of 2.45 GHz. Figure 3.11 (a) compares the S-parameters of the miniaturized FSS on the engineered substrate and the original FSS on a regular substrate, where similar performance is found. This means the engineered substrate helps to achieve a 16.02% size reduction without deteriorating the performance of the FSS. In addition, the operating frequency of the miniaturized FSS is tuned from 2.45 GHz to 2.672 GHz by applying DC current from 0 mA to 500 mA, as depicted in Figure 2.11 (b). The results fully demonstrate the design feasibility of miniaturized and tunable FSS on the engineered substrate optimized by the proposed model.

3.3.2 Miniaturized Antenna with Enhanced Performance

The increased permeability of the engineered substrate also brings great benefits to antenna designs, including better miniaturization, wider bandwidth, and higher efficiency. Figure 3.12 shows a simple patch antenna on the engineered substrate optimized by the proposed model. The antenna with a size of \(0.229\lambda_0 \times 0.302\lambda_0\) is first designed to operate at 2.45GHz on a normal FR4 substrate \((\varepsilon_r = 4.4, \mu_r = 1.0, \tan\delta = 0.02)\), which works as a reference. On the other hand, the optimized engineered substrate with 10 layers of permalloy thin films has an increased effective permeability of 2.398 for antenna
miniaturization. Thus, a miniaturized antenna (design 1) with the size of $0.171\lambda_0 \times 0.212\lambda_0$ is implemented on the engineered substrate, which means 47.58% size reduction. By comparing their reflection coefficients and radiation patterns in Figure 3.13 (a), it is concluded that the optimized engineered substrate can miniaturize the antenna without deteriorating its performance.

Another important parameter of an antenna is bandwidth. For a patch antenna, the zero-order bandwidth is approximated by [68]:

$$BW = \frac{96 \sqrt{\mu_r \epsilon_r} h}{\sqrt{2} \sqrt{4 + 17 \epsilon_r \mu_r}}$$  \hspace{1cm} (3.14)

where $h$ is the total thickness of the engineered substrate. Therefore, the antenna bandwidth is enhanced by increasing the ratio of $\mu_r / \epsilon_r$ for a given miniaturization factor ($\sqrt{\epsilon_r \mu_r}$). Compared to the original antenna on a regular dielectric substrate, the same miniaturization
factor is achieved by the antenna on the engineered substrate with lower permittivity.

Figure 3.13 (a) Performance comparison of miniaturized antenna (design 1) on the engineered substrate and original antenna on a regular dielectric substrate. (b) Performance comparison of the original antenna and design 2 with the same miniaturization factor.
leading to a wider bandwidth. Based on this, an optimized antenna (design 2) is implemented on the engineered substrate with the property of $\varepsilon_r = 1.835, \mu_r = 2.398$. As shown in Figure 3.13 (b), wider -10 dB bandwidth of 104.8 MHz is achieved by design 2, and a higher gain is found from the inserted radiation patterns. Compared with the 70 MHz bandwidth of the original antenna, a 49.71% bandwidth increase has been achieved. In addition, the increased permeability and decreased permittivity make the antenna impedance close to the air impedance, leading to a high radiation efficiency. The performance of the original antenna on a normal dielectric substrate and two optimized antennas on the engineered substrate are also compared in Table 3.2. In conclusion, the engineered substrate optimized by the proposed model can help an antenna achieve good miniaturization with increased bandwidth and improved radiation efficiency.

3.4 Conclusion

The engineered substrate with ferromagnetic thin films has found a lot of applications in designing miniaturized and tunable RF components with enhanced performance. In order to get higher effective permeability and a wider tuning range, multiple layers or thicker films are needed, which makes it critical to develop an accurate model. This dissertation presents a multilayer substrate model to characterize the

<table>
<thead>
<tr>
<th>Antenna</th>
<th>Size</th>
<th>$\sqrt{\varepsilon_r \mu_r}$</th>
<th>-10 dB bandwidth</th>
<th>Max gain</th>
</tr>
</thead>
<tbody>
<tr>
<td>Original antenna</td>
<td>$\lambda_0^2$</td>
<td>2.09</td>
<td>70MHz (2.86%)</td>
<td>3.5 dB</td>
</tr>
<tr>
<td>Design 1</td>
<td>$\lambda_0^2$</td>
<td>3.25</td>
<td>73MHz (2.98%)</td>
<td>3.4 dB</td>
</tr>
<tr>
<td>Design 2</td>
<td>$\lambda_0^2$</td>
<td>2.09</td>
<td>104.8MHz (4.28%)</td>
<td>4.2 dB</td>
</tr>
</tbody>
</table>

Table 3.2: Comparison of antenna performance
performance of the engineered substrate in terms of film thickness, number of film layers, film vertical position, and planar filling density. With the proposed model, the configuration of the engineered substrate can be quickly determined and optimized for the desired permeability. In addition, a miniaturized and tunable FSS and a performance enhanced antenna are implemented in this dissertation to demonstrate the design efficacy of the proposed model. Detailed guidelines for the design of RF components on the optimized engineered substrate are provided.
CHAPTER 4
HIGH-PERFORMANCE INTERCONNECTS WITH IMPROVED
SIGNAL INTEGRITY

4.1 Introduction

With the development of integrated circuits (ICs) and communication systems, there has been a progressively expanded tendency of increasing trace density on printed circuit boards (PCBs) and silicon dies in the semiconductor industry toward higher levels of integration and multi-core architecture generates chips with densities of hundreds of billions of transistors, driven by the requirement of low-cost and highly compact designs [69]. The market demands ever faster input/output signals with high density on compact boards, but the closer two parallel traces or conductors are placed together, the greater the electromagnetic field generated within one trace will interfere with the signal of the other, resulting in the near-end and far-end crosstalk (FEXT) noises. In general, FEXT is more critical than near-end crosstalk, because it will continue to increase along the trace, which seriously affects the signal integrity of the received signal at the receiving end. The resulting issues such as ground bounces, delays, and jitters will in turn deteriorate the performance of the high-speed and high-density system [70].
Multiple technologies have been applied to reduce the FEXT, among which, introducing short trapezoidal-shaped tabs between the coupled lines is the most common method [71]. However, there is still a limitation due to the surface tab routing itself and manufacturing capability. Technologies to further reduce the FEXT are highly desired for modern integrated systems with higher and higher data rates. In this section, the FEXTs of traditional coupled lines and the tabbed routing structure are first fully analyzed by extracting the corresponding capacitance and inductance matrices. Based on the analysis,
the FEXT can be further reduced by either increasing the mutual capacitance or increasing
the self-inductance. Thus, three improved designs are proposed and implemented for
mitigated FEXT. The signal integrity of high-speed and high-density interconnects is
significantly improved by the efficient methods and superior designs provided in this
section.

4.2 Far-End Crosstalk (FEXT) Analysis

Crosstalk is the coupling of energy from one signal to another and occurs from the
interaction of electric and magnetic fields between transmission lines. The far-end crosstalk
(FEXT) is defined as the difference between the capacitive coupling and the inductive
coupling between two adjacent signal lines. Figure 4.1 (a) shows the equivalent circuit of
a coupled microstrip line without considering the losses. If a signal is input to port 1, the
FEXT received at port 4 is calculated by [72]:

\[ FEXT = \frac{V_{in} l}{RT} \times \frac{1}{2v} \times \left( \frac{C_m}{C_s} - \frac{L_m}{L_s} \right) \]  (4.1)

where \( V_{in} \) is the input voltage, \( l \) is the line length, \( RT \) is the signal rise time, \( v \) is the speed
of the signal on the line, \( C_m \) is the mutual capacitance per length, \( C_s \) is the self-capacitance
per length, \( L_m \) is the mutual inductance per length, and \( L_s \) is the self-inductance per length.

To describe the inherent property of the coupled lines, a far-end coupling coefficient \( (k_f) \) is
defined as:

\[ k_f = \frac{1}{2v} \times \left( \frac{C_m}{C_s} - \frac{L_m}{L_s} \right) \]  (4.2)

Thus, it is necessary to extract the capacitance and inductance of the coupled lines under
quasi-static condition. Figure 4.2 (b) shows the capacitive components of the coupled
microstrip lines with arbitrary line widths. For each line, the self-capacitance \( C_{si} \) consists of parallel plate capacitance \( C_{pi} \), the external fringing capacitance \( C_f \), and the internal fringing capacitance \( C_{fe} \). Hence, the self-capacitance \( C_{si} \) are given as:

\[
C_{si} = C_{pi} + C_f + C_{fe} \quad i = 1, 2
\]  

(4.3)

The parallel plate capacitance \( C_{pi} \) are expressed as:

\[
C_{pi} = \varepsilon_0 \varepsilon_r \frac{w_i}{h} \quad i = 1, 2
\]  

(4.4)

The external fringing capacitance \( C_f \) and the internal fringing capacitance \( C_{fe} \) are approximately half of those in the coupled strip lines with substrate thickness of \( (2h+t) \) [73]. Thus, the external fringing capacitance \( C_f \) is given as:

\[
C_f = \frac{\varepsilon_0 \varepsilon_r}{\pi} \left[ 2t_b \ln(t_b + 1) - (t_b - 1) \ln(t_b^2 - 1) \right]
\]  

(4.5)

where \( t_b = (2h + t) / 2h \). The internal fringing capacitance \( C_{fe} \) is a function of the line thickness, the substrate height and the spacing between the lines. It is calculated by:

\[
C_{fe} = A \left\{ \frac{s}{2h+t} - \frac{2}{\pi} \ln \left[ \cosh \left( \frac{\pi}{2} \frac{s}{2h+t} \right) \right] \right\} + B
\]  

(4.6)

where \( A \) and \( B \) are defined as follows:

\[
A = \begin{cases} 
1.0 + 2.13507 \frac{t}{2h+t} \left( \frac{s}{2h+t} \right)^{-0.57518}, & \frac{s}{2h+t} < 0.3 \\
1.0 + 3.89531 \frac{t}{2h+t} \left( \frac{s}{2h+t} \right)^{-0.11467}, & \frac{s}{2h+t} \geq 0.3
\end{cases}
\]  

(4.7)

For \( t / (2h+t) \leq 0.5636 \),
\[ B = \begin{cases} 
- \left(0.2933 + 3.333 \frac{s}{2h+t}\right) \frac{t}{2h+t}, & \frac{s}{2h+t} < 0.08 \\
-0.56 \frac{t}{2h+t'}, & \frac{s}{2h+t} \geq 0.08 
\end{cases} \tag{4.8} \]

For \( t / (2h + t) > 0.5636 \),

\[ B = \begin{cases} 
-0.1653 - 5.6814 \frac{s}{2h+t} + 6.7475 \frac{t}{2h+t} \frac{s}{2h+t}, & \frac{s}{2h+t} < 0.08 \\
-0.62 + 0.54 \frac{t}{2h+t}, & \frac{s}{2h+t} \geq 0.08 
\end{cases} \tag{4.9} \]

The mutual capacitance \( C_m \) consists of the gap capacitance \( C_{m1} \) in the air above the coupled lines, the parallel-plate capacitance between the coupled lines \( C_{m2} \), and the gap capacitance \( C_{m3} \) in the substrate. Using these capacitances, the total mutual capacitance is written as:

\[ C_m = C_{m1} + C_{m2} + C_{m3} \tag{4.10} \]

The gap capacitance \( C_{m1} \) in the air above the coupled lines is calculated by [74]:

\[ C_{m1} = \varepsilon_0 \frac{K(k')}{K(k)} - \Delta C_f \tag{4.11} \]

\[ k = \frac{1+w_1/s+w_2/s}{(1+w_1/s)(1+w_2/s)} \tag{4.12} \]

\[ k'^2 = 1 - k^2 \tag{4.13} \]

where \( K(x) \) is the complete elliptical integral of the first kind. \( \Delta C_f \) is an empirical air-gap capacitance correction factor which is expressed by:

\[ \Delta C_f = \frac{(c_{f1}^{air} - c_{f1}^{irs})(c_{f2}^{air} - c_{f2}^{irs})}{c_{f1}^{air} + c_{f2}^{air} - c_{f1}^{irs} - c_{f2}^{irs}} \tag{4.14} \]
\[ C_{f_1}^{\text{air}} = 0.5 \left( \frac{1}{\varepsilon_0 Z_i} - \varepsilon_0 \frac{w_i}{h} \right), \quad i = 1, 2 \] (4.15)

\[ C_{f_{\text{sl}}}^{\text{air}} = 0.5 \left[ \frac{1}{\varepsilon_0 60\pi K(k_{\text{sl}})} - \varepsilon_0 \frac{w_i}{h} \right], \quad i = 1, 2 \] (4.16)

\[ k_{\text{sl}} = \tanh \left( \frac{\pi w_i}{4 h} \right), \quad i = 1, 2 \] (4.17)

\[ k_{\text{sl}}^{2} = 1 - k_{\text{sl}}^{2}, \quad i = 1, 2 \] (4.18)

where \( c_0 \) is the velocity of light in free space, \( Z_i \) is the characteristic impedance of single-ended microstrip line with air as substrate, and \( Z_i \) is given as:

\[
Z_i = \begin{cases} 
60 \ln \left( \frac{8h}{w_i} \right), & \frac{w_i}{h} \leq 1 \\
\frac{120\pi}{w_i + 1.393 + 0.677 \ln \left( \frac{w_i}{h} + 1.444 \right)}, & \frac{w_i}{h} \geq 1
\end{cases} \] (4.19)

The parallel-plate capacitance between the coupled lines \( C_{m2} \) is calculated by:

\[ C_{m2} = \varepsilon_0 \frac{t}{s} \] (4.20)

Finally, the gap capacitance \( C_{m3} \) in the substrate is expressed by:

\[ C_{m3} = \sqrt{C_{d1} C_{d2}} \] (4.21)

\[ C_{d_1} = 0.5 \left[ \frac{\varepsilon_r}{\varepsilon_0 60\pi K(k_{\text{ol}})} - \frac{\varepsilon_r}{\varepsilon_0 60\pi K(k_{\text{el}})} \right], \quad i = 1, 2 \] (4.22)

\[ k_{\text{ol}} = \tanh \left( \frac{\pi w_i}{4 h} \right) / \tanh \left( \frac{\pi s + w_i}{4 h} \right), \quad i = 1, 2 \] (4.23)

\[ k_{\text{ol}}^{2} = 1 - k_{\text{ol}}^{2}, \quad i = 1, 2 \] (4.24)

\[ k_{\text{el}} = \tanh \left( \frac{\pi w_i}{4 h} \right) \tanh \left( \frac{\pi s + w_i}{4 h} \right), \quad i = 1, 2 \] (4.25)
\[ k_{ei}^2 = 1 - k_{ei}^2, \ i = 1, 2 \]  

(4.26)

The self-inductance and mutual inductance are calculated from the self-capacitance and mutual capacitance when replacing the dielectric material by air [75]:

\[
\begin{bmatrix}
L_s & L_m \\
L_m & L_s
\end{bmatrix}
= \frac{1}{v_s^2} \cdot I_{2 \times 2} \cdot \begin{bmatrix}
C_s + C_m & -C_m \\
-C_m & C_s + C_m
\end{bmatrix}^{-1}
\]

(4.27)

where \( v_s \) is the wave velocity on the line within homogeneous media, \( I_{2 \times 2} \) is a 2\( \times \)2 identity matrix. It is worth pointing out that \( v_s \) is equal to the light speed in free space for uniform transmission lines with air as the substrate. The wave velocity needs to be corrected for non-uniform transmission lines, which will be introduced in the following.

To demonstrate the previous closed-form expressions of the capacitance and inductance, a model of coupled microstrip lines is simulated on a Rogers RO4350 substrate (\( \varepsilon_r = 3.66, \mu_r = 1, \tan \delta = 0.004 \)) by Ansys Q3D, an accurate quasi-static simulation tool. The dimensions of the coupled lines are arbitrarily chosen as: \( w_1 = w_2 = 1.08 \) mm, \( s = 1.2 \) mm, \( h = 0.5 \) mm, \( t = 35 \mu m, l = 40 \) mm. The results generated by the formulas and the simulation are compared in Table 4.1, where good agreement is found. The far-end coupling coefficient (\( k_f \)) is also calculated in the table to show the FEXT issue of the coupled microstrip lines. Methodologies are highly desirable to make the \( k_f \) close to zero, so that the FEXT is eliminated for high frequency applications.

<table>
<thead>
<tr>
<th></th>
<th>( C_m )</th>
<th>( C_s )</th>
<th>( L_m )</th>
<th>( L_s )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Q3D</td>
<td>0.10 pF</td>
<td>4.07 pF</td>
<td>0.86 nH</td>
<td>11.31 nH</td>
</tr>
<tr>
<td>Formula</td>
<td>0.11 pF</td>
<td>4.04 pF</td>
<td>0.91 nH</td>
<td>11.09 nH</td>
</tr>
</tbody>
</table>
4.3 Methodologies for FEXT Reduction

To improve the FEXT of coupled microstrip lines, interdigital trapezoidal tabs are added to increase the mutual capacitance, as shown in Figure 4.2 (a). In this dissertation,

Figure 4.2 (a) Tabbed routing structure. (b) Equivalent division of the unit segment. (c) Additional fringing capacitance to consider when calculating mutual capacitance.
such a tabbed routing structure is proposed for the first time to be analyzed by extracting the capacitance and inductance of the unit segment. Each unit segment is divided into five sections and these sections are approximately considered as coupled lines with different line widths, as illustrated in Figure 4.2 (b). With the previous closed-form formulas (4.3) and (4.10), the mutual capacitance and self-capacitance of the unit segment are calculated.

In addition, as depicted in Figure 4.2 (c), the coupling fringing capacitance between the tabs $C_{t1}$ and the fringing capacitance between the tab end and the lines $C_{t2}$ are additional capacitance components that need to be considered for the mutual capacitance calculation. The coupling fringing capacitance between the tabs $C_{t1}$ is calculated by considering it as a small section of coupled lines with length of $(2w_3 - s)$ and widths of $(l_1 + l_2) / 4$, $(l_1 + l_2) / 2$, respectively. The fringing capacitance is $C_{t2}$ given as:

$$C_{t2} = 0.5 \left( \frac{1}{\varepsilon_0 Z_t} - \varepsilon_0 \frac{l_1 + l_2}{2(s-w_3)} \right)$$

$$Z_t = \begin{cases} 
60 \ln \left( \frac{16(s-w_3)}{(l_1 + l_2)} + \frac{l_1 + l_2}{8(s-w_3)} \right), & \frac{l_1 + l_2}{2(s-w_3)} \leq 1 \\
\frac{l_1 + l_2}{120} + 1.393 + 0.677 \ln \left( \frac{l_1 + l_2}{2(s-w_3)} + 1.444 \right), & \frac{l_1 + l_2}{2(s-w_3)} \geq 1 
\end{cases}$$

Based on equation (4.27), the inductance matrix is calculated by the extracted capacitance matrix with air as substrate. It should be pointed out that the $v_s$ will not be equal to the light speed in free space because the integrated tabs make the original uniform line become a kind of slow-wave structure [76]. The wave velocity on the tabbed routing structure is then given as:

$$v_s = \frac{c_0}{SF}$$

where $SF$ is the slowing factor and is calculated by:
where $Z_a$ and $Z_b$ are characteristic impedances of the microstrip lines with the line widths of $w$ and $(w + w_t)$, respectively.

To verify the above analytical formulas, interdigital tabs are added to the original model of coupled lines in Section 4.2 to form a tabbed routing structure. The dimensions of the tabbed routing structure are chosen as: $w = 1.08$ mm, $w_t = 0.9$ mm, $l_1 = 0.52$ mm, $l_2 = 0.4$ mm, $l_3 = 0.68$ mm, $s = 1.2$ mm, $h = 0.5$ mm, $t = 35$ mm, $l = 40$ mm. The results generated by the formulas and the simulation are compared in Table 4.2, where good agreement is found. Compared to the original coupled lines, the mutual capacitance is significantly increased by the integrated tabs, resulting in the smaller far-end coupling coefficient ($k_f$).

<table>
<thead>
<tr>
<th></th>
<th>$C_m$</th>
<th>$C_s$</th>
<th>$L_m$</th>
<th>$L_d$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Q3D</td>
<td>0.33 pF</td>
<td>4.66 pF</td>
<td>0.95 nH</td>
<td>10.98 nH</td>
</tr>
<tr>
<td>Formula</td>
<td>0.34 pF</td>
<td>4.64 pF</td>
<td>1.04 nH</td>
<td>11.03 nH</td>
</tr>
</tbody>
</table>

Although the interdigital tabs can reduce the FEXT of coupled lines, it is usually not enough to fully eliminate the FEXT. To achieve further FEXT reduction, three improved structures are proposed in this dissertation. Based on equation (4.20), a straightforward way to increase the mutual capacitance is increasing the thickness of the tabs, as illustrated in Figure 4.3 (a). To minimize the impact on the inductance, the thickness of the main coupled lines is not changed so that the $k_f$ is closer to zero. In addition, the mutual capacitance can also be increased by increasing the permittivity ($\varepsilon_r$). Figure 4.3
(b) shows the second improved structure with a layer of high-permittivity dielectric.

Figure 4.3 Three improved structures to further reduce the FEXT: (a) 3D tabbed routing structure with thicker tabs, (b) tabbed routing structure with dielectric material between tabs, (c) tabbed routing structure with covered magnetic material.
between the two lines. The dielectric layer thickness is chosen as the same as the line thickness for controllable performance and easy fabrication. The dielectric layer will have no effect on the inductance, resulting in less FEXT. Instead of increasing the mutual capacitance, the FEXT can also be reduced by increasing the self-inductance of the two lines. Hence, as depicted in Figure 4.3 (c), a layer of magnetic thin films with high permeability ($\mu_r$) is proposed to cover on tabbed routing structure so that the self-inductance is significantly increased, which will also decrease the FEXT.

![Comparison of the simulated FEXT ($S_{41}$) of various interconnects.](image)

Figure 4.4 Comparison of the simulated FEXT ($S_{41}$) of various interconnects.

$S$-parameters ($S_{41}$) are simulated to evaluate the FEXT performance of different cases. The proposed structures use the same dimensions and same substrate as the prementioned tabbed routing structure. For the tabbed routing structure with thicker tabs, the increased tab thickness is chosen as $t_1 = 140 \, \mu m$. For the tabbed routing structure with dielectric material, the permittivity of the dielectric is chosen as $\varepsilon_r = 4$. For the last structure
with covered magnetic material, the thin film thickness $t_2 = 35 \, \mu m$ and the permeability is set as $\mu_r = 2000$. Figure 4.4 gives the simulated FEXT of those cases. The tabbed routing structure is simulated to have less FEXT than the original coupled lines. In addition, the proposed three structures can further improve the FEXT as expected. Either way can get
better FEXT reduction by changing the corresponding parameters to minimize the absolute value of $k_f$.

|                         | $C_m$  | $C_s$  | $L_m$  | $L_s$  | $|k_f|$   |
|-------------------------|--------|--------|--------|--------|-----------|
| Coupled lines           | 0.09 pF| 7.25 pF| 1.42 nH| 19.21 nH| 28.69×10⁻¹¹ |
| Tabbed routing          | 0.37 pF| 8.82 pF| 1.49 nH| 18.77 nH| 19.04×10⁻¹¹ |
| Thicker tabs            | 0.43 pF| 8.75 pF| 1.50 nH| 18.75 nH| 15.62×10⁻¹¹ |
| Add dielectric material | 0.88 pF| 8.82 pF| 1.49 nH| 18.77 nH| 9.79×10⁻¹¹  |
| Cover magnetic material | 0.37 pF| 8.82 pF| 1.21 nH| 21.61 nH| 7.60×10⁻¹¹  |

4.4 Fabrication and Measurement

To verify the performance of the proposed structures, all the aforementioned structures are fabricated on the silicon wafer ($\varepsilon_r = 11.9$, $\mu_r = 1$, tan $\delta = 0.005$). The dimensions are chosen as: $t = 100$ nm, $t_1 = 400$ nm, $t_2 = 100$ nm, $w = 0.39$ mm, $w_t = 0.7$

Figure 4.6 Schematic diagram of the fabrication process flow. Steps (3a), (3b), and (3c) were performed on separate wafers to independently assess the impact on device performance.
mm, $s = 1.2 \text{ mm}$, $l = 40 \text{ mm}$, $l_1 = 0.52 \text{ mm}$, $l_2 = 0.4 \text{ mm}$, $l_3 = 0.68 \text{ mm}$, $h = 0.5 \text{ mm}$.

Ferroelectric material Barium strontium titanate (BST, $\varepsilon_r = 600$, $\mu_r = 1$) is deposited

![Figure 4.7 Photographs of the fabricated devices: (a) coupled lines, (b) tabbed routing structure, (c) tabbed routing structure with thicker tabs, (d) tabbed routing structure with BST, (e) tabbed routing structure with permalloy.](image)
between the tabs, and the ferromagnetic material permalloy ($\varepsilon_r = 1$, $\mu_r = 1000$) is chosen to cover the metal lines. The calculated capacitance, inductance and the absolute values of $k_f$ are listed in Table 4.3. It is noted that the original coupled lines have the highest $k_f$ while the tabbed routing structure can improve it to a certain degree. In addition, the proposed three structures can further reduce the $k_f$, resulting in smaller FEXT. Figure 4.5 (a) shows the specially designed circuit designed for the on-wafer probe testing for those devices under test (DUT). Microstrip line to coplanar waveguide (CPW) transition is optimized for the ground-signal-ground (GSG) probe connection, as illustrated in Figure 4.5 (b).

Those samples are fabricated by photolithography technology, and a schematic diagram of the process flow is shown in Figure 4.6. The device fabrication procedure consists of first using a modified Radio Corporation of America (RCA) clean to prepare the wafer. Then, the wafer is patterned with photoresist and windows are opened for electron beam (e-beam) deposition of a Ti (30 nm) / Au (100 nm) metal stack to form the coupled lines with and without tabs. For the tabbed routing structure with thicker tabs, a second photoresist patterning followed by the deposition of an additional 300 nm Au layer is performed such that only the tabs are thickened. For the tabbed routing structure with a BST layer, the coupled lines and tabs are patterned followed by e-beam deposition of 100 nm thick BST. The final tabbed routing structure with a 100 nm permalloy cap layer is similarly formed by photoresist patterning and e-beam deposition. In all cases, an acetone-based lift-off procedure is used after the e-beam deposition followed by a heated organic cleaning using Acetone and 2-Proponal. In addition, all the devices have a fully deposited metal layer on the backside that serves as a ground. The photographs of the fabricated samples are shown in Figure 4.7.
The measurements of these samples are conducted after the standard on-wafer

Figure 4.8 Comparison of the measured and simulated (a) FEXT ($S_{41}$) and (b) insertion loss($S_{21}$).
Short-Open-Load-Through (SOLT) calibration with the CS-5 calibration substrate. The measured and simulated results are compared in Figure 4.8, where good agreement is found. It is found from Figure 4.8 (a) that the tabbed routing structure can decrease the FEXT of the original coupled lines to a certain degree. The FEXT becomes smaller for the tabbed routing structure with thicker tabs. The limited reduction in FEXT is due to the limited increase in the mutual capacitance of the 300 nm thicker tabs. Better performance is achieved by further increasing the tab thickness. In addition, the tabbed routing structures with BST and permalloy both have significant improvements in the FEXT as predicted in Table 4.3. Table 4.4 gives the specific measured FEXT ($S_{41}$) values versus frequencies for the above samples, where advantages of the proposed designs are obvious. The measured

Table 4.4: Specific measured FEXT ($S_{41}$) values versus frequencies for different structures (dB)

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>6</th>
<th>8</th>
<th>10</th>
<th>12</th>
<th>14</th>
</tr>
</thead>
<tbody>
<tr>
<td>Coupled lines</td>
<td>-15.60</td>
<td>-12.48</td>
<td>-10.99</td>
<td>-10.34</td>
<td>-9.50</td>
</tr>
<tr>
<td>Tabbed routing</td>
<td>-24.63</td>
<td>-18.61</td>
<td>-15.00</td>
<td>-14.22</td>
<td>-14.73</td>
</tr>
<tr>
<td>Thicker tabs</td>
<td>-27.33</td>
<td>-21.00</td>
<td>-17.44</td>
<td>-16.65</td>
<td>-16.83</td>
</tr>
<tr>
<td>Add dielectric material</td>
<td>-34.76</td>
<td>-28.57</td>
<td>-25.19</td>
<td>-23.94</td>
<td>-25.65</td>
</tr>
</tbody>
</table>

Figure 4.9 Circuits for eye diagram simulations in ADS based on the measured SNP files.
results successfully demonstrate the performance of the proposed designs. Figure 4.8 (b) also compares the insertion loss of different cases. The high insertion loss of these samples is mainly caused by the high resistance of the very thin metal layer. In addition, the high FEXT of coupled lines usually leads to even worse insertion loss since part of the energy

![Figure 4.8 (b) and Figure 4.10 Eye diagrams](image)

Figure 4.10 Eye diagrams of (a) coupled lines, (b) tabbed routing structure, (c) tabbed routing structure with thicker tabs, (d) tabbed routing structure with BST, (e) tabbed routing structure with permalloy.
is delivered to port 4, which can be significantly improved by the proposed methods. Compared to the original coupled lines, the tabbed routing structure and the tabbed routing structure with thicker tabs both have lower insertion loss. Similarly, the tabbed routing structure with BST has even better insertion loss due to the further improvement of FEXT. However, the tabbed routing structure with permalloy does not have a significantly improved insertion loss. This is mainly caused by the relatively low conductivity of the permalloy, which introduces more loss to the transmission lines. There is a tradeoff between the FEXT and insertion loss. Nevertheless, it is still necessary to introduce this permalloy based FEXT reduction method, because it provides a good solution to migrate the negative effects on the transmission line impedance caused by the other FEXT reduction methods. Properly integrating the permalloy and BST thin films should be a good way to achieve significant FEXT reduction, acceptable insertion loss, and constant characteristic impedance.

To evaluate the high-speed data transmission performance of the above samples, the measured SNP files are imported to the Advanced Design System (ADS) software to simulate the corresponding eye diagrams, as illustrated in Figure 4.9. A randomly selected 12-Gb/s pseudorandom binary sequence (PRBS) with 50 psec rise/fall time is applied to the transmitter (Tx) port (port 3). The crosstalk (Xtlk) from port one is set to have the same signal swing at the transmitter but random relative phase. Subsequently, the influences of the asynchronous crosstalk on channel performance are obtained from the eye diagrams at receiver (Rx) port (port 4). Figure 4.10 compares the eye diagrams of the above five samples, where a similar trend to measured S-parameters is found. The eye height and eye width of the coupled lines are 0.09 V and 47 psec, respectively, resulting in a small eye
opening. The eye opening is improved by the tabbed routing structure due to the reduced FEXT. Moreover, much better eye diagrams are achieved by the proposed three structures.

4.5 Conclusion

In this section, the FEXT of traditional coupled lines and tabbed routing structure are fully analyzed by extracting closed-form formulas of the corresponding capacitance and inductance matrices. Methods of reducing the FEXT are discussed, and three designs are proposed to further reduce the FEXT with 3D configurations and implementation of ferromagnetic and ferroelectric thin films. Photolithography technology is employed to fabricate the proposed designs and the measured results fully demonstrate the efficacy of the proposed methods and validate the generated formulas. This dissertation provides effective and flexible solutions to improve the signal integrity of high-speed and high-density systems.
CHAPTER 5
SUMMARY AND FUTURE WORK

5.1 Summary of Contributions

In this dissertation, novel structure designs and thin film techniques are employed to design multi-functional and reconfigurable RF components, which enables simplified and miniaturized wireless communication systems. The thin film techniques are also applied to solve the associated signal integrity issue of the highly integrated systems.

The first part of the dissertation proposes a filtering structure consisting of two half-wavelength resonators and two open-stub loaded resonators. With the novel coupling scheme, two third-order passbands with high selectivity and good out-of-band rejection are achieved. In order to achieve switchable filtering responses, stepped-impedance open stubs and one-end-shorted coupled lines are employed to suppress unwanted signal. By controlling the operating states of the signal suppression methods, four-state filtering responses are achieved: both passbands ON, both passbands OFF, high-frequency passband ON, and low-frequency passband ON. In addition, the filtering structure is also employed to design a dual-functional filtering balun by integrating the double-sides parallel-strip lines (DSPSL). DSPSL is placed at the port of two identical filtering structure layers separated by an intermediate ground layer. Due to the inherent out-of-phase feature of DSPSL, the proposed filtering balun can convert the input unbalanced signal into two balanced signals while maintaining the dual-band filtering function. The measured
results of both the switchable filter and the filtering balun show great agreement with the simulated results, which demonstrates the design efficacy of the proposed filtering structure. The proposed mutual-functional filter designs can significantly simplify and miniaturize the communication systems, showing the great benefits of the multifunctional RF technologies.

The second part explores more solutions for miniaturized systems based on the engineered substrate with tunable permeability. Ferromagnetic materials, such as permalloy, have high permeability and the permeability can be tuned by external magnetic field. By integrating ferromagnetic thin films into a normal RF substrate, an engineered substrate with increased effective permeability is enabled. The permeability is tuned by applying different DC currents to the metal bias lines beneath the ferromagnetic thin films. The electrical tuning mechanism and strategies for improving the FMR frequency and magnetic loss are discussed in detail. In order to obtain higher effective permeability and wider tuning range of the engineered substrate, more layers of ferromagnetic thin films are required, making it necessary to create an accurate model for configuration optimization and performance prediction. In this section, microstrip line model is employed to thoroughly explore the performance of the engineered substrate on the effects of number of film layers, film thickness, film vertical locations, planar filling density, and pattern dimension. Conformal mapping theory is applied to quickly determine the effective area of ferromagnetic thin films for various microstrip lines. The effective permeability of the engineered substrate is characterized by simulating the inductance of the microstrip line model under different scenarios. To simplify the characterization process, the relationship between the inductance and the substrate permeability is also derived and expressed by
closed-form formulas. Finally, the engineered substate model is developed and verified with numerous simulation cases. To demonstrate the design efficacy of the proposed engineered substrate model, a miniaturized frequency selective surface and a performance enhanced patch antenna are implemented on the engineered substrate optimized by the model. In addition, the increased permeability of the engineered substrate can significantly improve the bandwidth of an antenna, which makes it very promising for antenna designs, especially for ultra-wide band antennas.

With the help of multi-functional and reconfigurable RF technologies, a miniaturized communication system is achieved. However, the crosstalk between the high-density traces in the highly integrated system will significantly deteriorate the system signal integrity. In the final part of this dissertation, the far-end crosstalk of coupled lines and the tabbed routing structure are analyzed comprehensively. Closed-form formulas are derived to directly calculate the capacitance and inductance matrices, thereby determining the FEXT. Based on the theoretical analysis, three improved methodologies are proposed to further eliminate the FEXT, i.e., increasing the tab thickness and keeping the main lines unchanged, introducing high permittivity ferroelectric thin films to the selected area (gap) between the tabs, and covering all the metal lines with high permeability ferromagnetic thin films. Those design samples are deposited on the silicon wafers by photolithography technology. The measured results show the further improved performance of the proposed methodologies.
5.2 Future Work

Based on the summary of finished work, the future work will focus on exploring more functions and applications by novel structure designs or integrating RF components with thin film techniques. The proposed filtering structure in Chapter 2 can be used to design filters with more functions and better filtering performance. For example, more passbands can be generated by introducing more transmission paths with more resonators [77]. By properly integrating resistors, a filter absorber can also be achieved with both filtering function and signal absorbing capability [78]. In addition, the multi-functional filters can be placed on the engineered substrate presented in Chapter 2, which makes it possible to achieve better miniaturization and tunable operating frequency [79]. The thin film enabled engineered substrate in Chapter 2 can be further improved by follows: (1) improving the tuning efficiency and tunability range by investigating the configurations and types of ferromagnetic thin films; (2) reducing magnetic loss by utilizing various magnetic materials (e.g., ferrite) with high resistivity, which can help to eliminate unwanted eddy current; (3) integrating both ferromagnetic and ferroelectric materials inside a design to realize dual inductive and capacitive tunability; and (4) developing the integration methodology of ferromagnetic materials for a wider range of microwave components. The thin film techniques are also employed to decrease the FEXT between coupled lines in this dissertation. In the future, ferromagnetic thin films and ferroelectric thin films can be integrated together on the tabbed routing structure to achieve even better FEXT reduction. In addition, the negative effects of the additional tabs between should be investigated, including insertion loss and phase delay. This leads to another potential advantage of the thin film techniques: they can help the tabbed routing structure achieve
the same low FEXT but with smaller tabs, which minimizes impedance mismatches caused by these tabs.
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APPENDIX A
MULTIPORT VNA MEASUREMENT

Vector network analyzer (VNA) calibration is of great importance for precise measurement of the device under test (DUT). Cables, connectors, the test fixture and the like all systematic errors in the system and they must be eliminated before the VNA is used so that correct measurements can be made. Currently, calibration methods such as Short-Open-Load-Through (SOLT) [80], Through-Reflection-Line (TRL) [81], and Through-Open-Match (TOM) [82] are widely used for VNA measurements. These methods usually require standard calibration kits or calibration substrates to provide an ideal short circuit, an open circuit, a through connection, and a precision load (usually 50 Ω). The switchable filter in Chapter 1, the filtering balun in Chapter 1, and the samples in Chapter 4 need two-port, three-port, and four-port VNA measurements, respectively. Hence, the multiport VNA calibration and measurement process are provided in this appendix.

R&S ZVA 67 vector network analyzer and SOLT calibration method are used to measure the fabricated samples in this dissertation. For the filter measurement in Chapter 1, the Agilent 85052D 3.5 mm calibration kit is used for the calibration with standard SMA connectors. For the on-wafer measurement in Chapter 4, calibration substrate CS-5 is used for the calibration with 40-A-GSG-200-DP probes. Generally, the VNA measurement process is shown as follows: (1) Set the frequency range of measurement with start/stop frequency or center frequency/span. (2) Set the frequency step size (number of points). (3)
Set the intermediate frequency (IF) bandwidth. Smaller IF bandwidth has lower noise floor and higher accuracy, while longer sweeping time. (4) Select and set the calibration kits of short, open, load, and through, including the number of standards, circuit model, and frequency range. In this dissertation, 2.92 mm cables are set and used to measure signals at frequency as high as 44GHz. (5) Set the SOLT calibration with the desired number of ports. Connect each port to the standard calibration kits or substrates and calibrate the corresponding short, open, load, and through. It is worth mentioning that the through calibration needs to be done for every combination of two ports. For example, six combinations of ports (e.g., port 1&2, port 1&3, port 1&4, port 2&3, port 2&4, port 3&4) need to be calibrated for the four-port measurement. (6) After the calibration is accepted, use the through standard and the Smith Chart to check the qualification of calibration. A perfect calibration should put the reflection coefficient at the center (perfectly matched) of Smith Chart if two ports are connected with a through standard. (7) Measure the DUT.