The Field Programmable Microwave Substrate

by

Nathan Jess, B.Eng., M.A.Sc.

A thesis submitted to the
Faculty of Graduate and Postdoctoral Affairs
in partial fulfillment of the requirements for the degree of

Doctor of Philosophy in Electrical Engineering

Ottawa-Carleton Institute for Electrical and Computer Engineering

Department of Electronics

Carleton University

Ottawa, Ontario

January, 2016

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Abstract

This thesis describes a research project to create a programmable microwave circuit having a similar level of programmability as that of a field programmable gate array (FPGA). The result of the thesis is the realization of the first ever practical low-loss programmable microwave waveguides. Since waveguides are a key component in any microwave circuit, this allows for the realization of a broad range of microwave functions with a single circuit. The programmable waveguides are implemented in what is referred to as a field programmable microwave substrate. The programmable substrate is implemented using metamaterials between two parallel metal plates. Between the metal plates are electrically small unit cells that consist of metal structuring that connects with active components not contained between the metal plates (i.e. above or below the metal plates). Each of these unit cells can be programmed to have a range of positive dielectric constants or a negative dielectric constant. Programming a positive dielectric constant core and negative dielectric constant sidewall yields a structure that can be described using the slab waveguide equations. This allows for waveguides to be dynamically programmed within the field programmable microwave substrate.

The theory required to design low-loss metamaterials is presented and used to realize two prototype circuits. The first implementation is on an FR4 substrate and is used to demonstrate low-loss waveguides, amplifiers and oscillators from 0.9 –

3GHz. The other field programmable microwave substrate is implemented in low temperature co-fired ceramic and uses custom designed CMOS chips. This version is used to demonstrate miniaturization. The losses of the waveguides are high, but this is mainly due to manufacturing capabilities of the process used. Low-loss small field programmable microwave substrates should be possible to implement with the use of higher performance components and more sophisticated manufacturing processes.

Acknowledgments

First, I would like to thank my thesis supervisors Barry Syrett and Langis Roy for their support and allowing me the freedom to pursue my own interests and curiosities. I would also like to thank the support staff in the department of electronics; notably, Nagui Mikhail, Scott Bruce, Blazenka Power, Anna Lee, and Sylvie Beekmans. I would also like to thank Professor Steve McGarry for many practical conversations and assistance. There are numerous fellow grad students that I would like to thank. Tyler Ross and Che Knisely were excellent for chatting with on technical issues and great for helping solve problems. Thomas Pepler provided a significant amount of assistance implementing the digital CMOS circuit for one of my designs. Ryan Griffin helped debug some circuits during testing. A general thanks goes out to the graduate students in the blue room who have made doing a PhD a pleasant experience. Finally, I would like to thank my family and wonderful girlfriend Jessica Lynch for all their support over the years.

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List of Abbreviations

ADC	analog-to-digital converter
ASIC	application specific integrated circuit
CEW	continuous electrowetting
CMOS	complementary metal-oxide-semiconductor
CPW	coplanar waveguide
DAC	digital-to-analog converter
ECC	electrochemically controlled capillarity
EM	electromagnetic
ESD	electrostatic discharge
EWOD	electrowetting on dielectric
FPGA	field programmable gate array
FPMC	field programmable microwave circuit
FPMS	field programmable microwave substrate

HEMT	high-electron-mobility transistor
LTCC	low temperature co-fired ceramic
MEMS	microelectromechanical systems
NRW	Nicolson-Ross-Weir
PCB	printed circuit board
PCM	phase change material
PEC	perfect electric conductor
PMC	perfect magnetic conductor
PROMFA	programmable microwave function array
RF	radio frequency
SDR	software defined radio
SIW	substrate integrated waveguide
SLCFET	super-lattice castellated field effect transistor
SOI	silicon on insulator
SOS	silicon on sapphire
VCO	voltage controlled oscillator
VNA	vector network analyzer

Chapter 1

Introduction

1.1 Time for Wireless to Change

Wireless technology has become ubiquitous with modern society and can be found in many electronic devices. Frequently these devices contain more than one radio, which are currently realized using a separate circuit for each frequency band and communication standard [1]. As the number of frequency bands and standards required in a device increases, this approach will become too costly and will likely lead to another solution being adopted. The ideal solution would be a radio with hardware that could be dynamically reconfigured for use over a wide range of frequencies and be able to support any communications standard. Another method that has received a significant amount of research is known as the software defined radio (SDR). This approach attempts to digitize the signal as close to the antenna as possible, preferably at the antenna as shown in Figure 1.1, and do all the signal processing in software rather than in hardware [2]. Unfortunately, the requirements on the digital-to-analog and analog-to-digital converters are beyond current and near future technical capabilities [3]. For this reason the current implementations involve the use of reconfigurable and programmable hardware in conjunction with software [3]. Both

Lime Microsystems [4] and Analog Devices [5] produce a integrated circuit that can be used for SDR applications. These approaches are based on radio frequency integrated circuit design techniques and this circuit design approach is less than optimal at high frequencies.

Figure 1.1: Mitola's concept of an ideal SDR [2].

Field programmable microwave circuits (FPMC) are currently a nascent field. FPMCs are envisioned to have the ability to be reprogrammed with the same level of reconfigurability as the field programmable gate array (FPGA). The FPGA was originally introduced to reduce reliance on digital application specific integrated circuits (ASICs). The idea is that a general purpose digital circuit can be used to significantly reduce non-recurring engineering costs and eliminate much of the risk associated with ASICs. Since the introduction of the FPGA in the 1980's it has flourished and become commonplace in many applications such as with hobbyists and in telecommunications [6]. Microwave communication systems suffer from the same high cost, high risk problems as early digital circuits, but implementation of FPMCs remains elusive [7].

1.2 Field Programmable Microwave Circuits

Design of a general purpose platform requires one to first answer a question. What are the basic building blocks? The waveguide is among the most basic of components in microwave circuits. Given a waveguide and an active device with gain, many of the microwave circuits required for a communications link can be realized. This dissertation presents a structure that is referred to as a field programmable microwave substrate (FPMS). This is a programmable substrate that allows for the material properties to be dynamically programmed with a certain resolution. The material properties can be programmed to have a range of positive dielectric constants or switched to a highly negative dielectric constant. This is achieved with programmable artificial dielectrics, or metamaterials. A waveguide is realized using a positive dielectric constant material sandwiched between negative dielectric constant materials. These waveguides are then used to realize components such as splitters/combiners, filters and by adding active devices, amplifiers and oscillators.

1.3 Thesis Overview

This work presents the field programmable microwave substrate (FPMS). Chapter 2 contains a review of the literature required to realize an FPMS. This includes a review of the components used to realize reconfigurable microwave circuits. The focus of the review is on switch technologies, liquid metals, and varactors. Components that facilitate reconfigurable behavior and are not considered as important to this work are only reviewed briefly. This is followed by an overview of both double positive metamaterials and double negative metamaterials. In addition, reconfigurable metamaterials are quickly reviewed. Next, a review of highly reconfigurable and general purpose microwave circuits is given. The chapter is concluded with an overview

of substrate integrated waveguide (SIW) circuits. The SIW circuits are important because they function in a similar manner to the programmable waveguides presented in this dissertation.

Chapter 3 analyzes an artificial dielectric using a transmission line equivalent circuit of a parallel plate waveguide. The goal of this chapter is to give the reader an understanding of design choices for designing both positive and negative artificial dielectrics. The primary focus is on methods to minimize the effects of lossy components. In general, programmable behavior requires the use of a switch and the losses from this switch need to be minimized. The equivalent circuit is simulated in ADS and the s-parameters are used with the Nicolson-Ross-Weir (NRW) algorithm to extract the effective material properties. A positive artificial dielectric is analyzed first with the equivalent circuit and a position to place a lossy component is identified. This position allows for the losses from the switch to be minimized. Next, various circuit elements are changed to find methods to minimize losses from the lossy component further. This is followed by a similar analysis for a negative artificial dielectric. The chapter is concluded with a brief analysis of waveguides constructed using the materials previously analyzed.

Chapter 4 presents two different implementations of the FPMS. One FPMS is implemented using an FR4 substrate. This implementation is able to function from $0.9 - 3GHz$ and is used to demonstrate programmable waveguides, amplifiers, and oscillators. A miniaturized version of the FPMS is presented next. This version is fabricated in LTCC and uses custom designed CMOS chips flip-chip bonded to the LTCC tile. This implementation is used to demonstrate waveguides from $2.4 - 3.4GHz$.

Chapter 5 is a discussion on various aspects on the FPMS. This includes various thoughts on the FPMS technology such as limitations, improvements, and potential

future work.

Chapter 6 gives a brief conclusion, presents limitations and implications for the FPMS, and lists contributions.

Chapter 2

Programmable Communications Hardware

2.1 Introduction

Currently, programmable communication technology is an emerging field. The field is coming into existence due to a need for ever more functional communications hardware. Radios are being attached to an ever increasing number of devices and many devices are required to function over a broad range of frequencies and standards. The growing requirements for communications hardware has led to significant research in advancing communications technology. Most of the advancements in communications hardware have been in the form of circuit performance and miniaturization. Currently, there is more demand for microwave circuits capable of functioning over a large range of frequencies and waveforms. This can be accomplished with a combination of hardware and software or primarily in hardware. This chapter focuses on the hardware used to realize reconfigurable microwave circuits. The next section reviews materials and components used to implement tunable and reconfigurable circuits. The following section discusses metamaterials and methods to realize reconfigurable and tunable metamaterials. Next, highly programmable microwave and RF circuits are presented. This is followed by a presentation of substrate integrated waveguide

circuits with significant levels of reconfigurability.

2.2 Reconfigurable and Tunable Materials and Components

There are a large number of materials and components that have been developed over the years to realize reconfigurable and tunable microwave circuits. Figure 2.1 shows a breakdown of most of the methods used to make circuits tunable or reconfigurable. Reconfigurable circuits can be realized using semiconductors, materials with reconfigurable properties, or by mechanically changing the circuit. Semiconductor switches and varactors are the most commonly seen components used to realize reconfigurability because they can be cheaply integrated on the same chip with the RF/microwave circuitry. Many of the other methods are still in development or are only seen in applications that require better performance and can justify the cost increase. It should be noted that a MEMS antenna tuning module produced by WiSpry is used in a mass-produced smartphone [8] and MEMS switches and varactors are likely to become more common in consumer products in the near future [9]. This section discusses each of the methods to achieve reconfigurability. In general, methods that obtain switching behaviour or extremely large changes in capacitance are considered important and are discussed in depth, while methods that are seen as less useful for this work will be presented briefly.

Figure 2.1: Methods used to make reconfigurable circuits. *Utilizes one of the other methods to create reconfigurable circuits.

2.2.1 Mechanical Tuning

Mechanically changing a microwave circuit is a useful tool since microwave circuit performance is intimately linked with physical dimensions. Physically changing the circuit can yield extremely large range of reconfigurability, but it is slow and can be prone to problems due to wear and tear. Broadly, mechanical means of tuning can be broken down to MEMS, mechanical deformation, microfluidics, and memristors.

MEMS Devices

MEMS allow for microwave circuits to be reconfigured through actuation of generally metallic structures. The majority of MEMS devices are actuated using an electrical bias, however some devices use electro-thermal activation. One such device uses an inter-digitated beam for capacitance where one beam is movable using electro-thermal actuators to change the gap [10]. Characteristics of MEMS switches and varactors are [11]:

- Excellent performance up to 100GHz.
- Low loss (<0.1 - 0.2 dB) and low on-resistance ($0.5 - 2\Omega$).
- Low off-state capacitance (2-16fF) up to high frequencies.
- Negligible power consumption and only during switching.
- High power handling.
- Capacitive switches have ratios of 20-150:1.
- Varactors have tuning ratios of 1.5-8:1.
- Switches have a figure of merit $R_{ON} \times C_{OFF}$ of 2.5fs [12].
- Large bias voltages required (25-90V).
- Hermetic packaging required.
- Some reliability issues.

MEMS devices such as switches [13–15] and varactors [16, 17] are fabricated on a variety of materials including silicon, gallium arsenide, silicon dioxide, LTCC, and ceramic.

MEMS switches [18–22] and varactors [10, 23] are used in a variety of reconfigurable microwave circuits due to their low loss and high power handling capability. MEMS are normally used to replace semiconductor switches and varactors in reconfigurable microwave circuits. However, they can be used in ways that semiconductor components cannot. One such structure that realizes significant tunability is the capacitive-post loaded resonant cavity shown in Figure 2.2 [24]. The resonant frequency of the cavity can be tuned from 6.1-24.4GHz. This is done by actuating the gold diaphragm from $0.5 - 15\mu m$ above the post.

Figure 2.2: Tunable capacitive-post loaded resonant cavity. (a) is the cylindrical cavity and (b) is a close up of the MEMS section. In (b) there is a thin insulator (red) above the post (orange), then a thin layer of gold (yellow) on silicon (grey) for the diaphragm. There is then a gap for actuation of the diaphragm. After the gap is a dielectric layer with a layer of silicon on top for an electrode [24].

Mechanical Deformation

There are numerous possible ways to tune a microwave circuit by mechanically changing the circuit. A common method of tuning microwave components is with the use of a tuning screw [25]. A more unique method of tuning a microwave circuit with mechanical deformation is with stretchable electrically conductive materials. These

devices are fabricated on substrates of stretchable polymers such as Polydimethylsiloxane (PDMS) and the conductor is an electrically conductive composite of silver flakes loaded in a platinum cured PDMS. The frequency response of these structures can then be lowered by physically stretching the circuit. The filter presented in one paper shows a shift of the cut-off frequency for the low-pass response from 1.4-1.2GHz with a 40% elongation. The insertion loss of the filter at 500MHz is 2.3dB [26].

Microfluidics Tuning

Microfluidics involves pumping either a liquid dielectric or conductive material into or out of a section of a microwave circuit. One paper reports a tunable dielectric resonant antenna that is reconfigurable using a liquid dielectric [27]. The tunability is realized with a solution of hydrotreated naphthenic oil with colloidal barium strontium titanate dispersed in it. This solution is placed in a cylindrical polycarbonate shell on a ground plane excited by a coaxial probe through the ground plane. Varying the height of the liquid from 8-24mm results in a resonant frequency change from 4.22-2.75GHz.

Microfluidics for microwave circuits are more commonly used for moving around a liquid metal. The recent development of a non-toxic liquid metal known as Galinstan has led to significant investigation into reconfigurable liquid metal microwave circuits. Galinstan is reported to have the following characteristics [28–30]:

- Eutectic consisting of gallium, indium, and tin.
- Conductivity of $2.3 \times 10^6 S/m$.
- Liquid from -19 to $1300^\circ C$.

- Oxidizes in as little as 1ppm of oxygen. This oxide can be reduced in an electrolyte solution such as HCl or NaOH.
- Easily alloys with many metals.
- Has a high surface tension and tends to ball up.

Galinstan can be moved in four different ways: fluidic pumping [31]; electrowetting on dielectric (EWOD) [29]; continuous electrowetting (CEW) [30]; and electrochemically controlled capillarity (ECC) [32].

Fluidic pumping involves using another liquid to push the metal to a different location. Fluidic pumping can be used to realize a capacitive switch as shown in Figure 2.3. The switching is achieved by moving the liquid metal on top of a coplanar waveguide (CPW) using a Teflon solution that is pumped in to push the liquid metal. The switch has an insertion loss from 0.1dB-0.47dB for a frequency range of 2 – 100GHz [33]. Another circuit implemented by pumping Galinstan is a reconfigurable substrate integrated waveguide filter. The filter is reconfigured using four hollow vias that can be filled or emptied using a syringe. After emptying, the via is flushed with an NaOH solution to dissolve any residual oxidized Galinstan. The filter can be reconfigured from 10.13-11.31GHz, has an insertion loss of 1.35dB, and an unloaded quality factor of 154 [34].

Figure 2.3: Capacitive liquid metal switch implemented with CPW [33].

Figure 2.4: Electrowetting on dielectric. Unbiased liquid metal shown in (a) and biased liquid metal shown in (b) [29].

Electrowetting on a dielectric involves modifying the effective surface energy of the liquid metal by applying a voltage differential as shown in Figure 2.4 [29]. One circuit reported using the EWOD method is a 12-18GHz electrically tunable resonant cavity as shown in Figure 2.5 which has a quality factor of 1400-1840 [29]. This circuit uses a drop of Galinstan to realize a change in capacitance with applied voltage, similar to a varactor. One gold layer connects to the Galinstan drop while the other gold layer is insulated from the drop with a layer of Teflon. Applying a bias from 0-100V across the two gold layers causes the Galinstan drop to deform and spread out along the Teflon. The deformation is proportional to the applied voltage and the drop becomes spherical again with no voltage applied. This results in a tunable capacitance between the post and the cavity wall through the Galinstan drop. The paper

reports that the Galinstan drop locks very robustly to the piece of metal connecting to the Galinstan drop.

Figure 2.5: Liquid metal tunable capacitive-post loaded cylindrical resonant cavity. Full structure shown in (a) and zoomed in view of liquid metal shown in (b) [29].

CEW involves immersing the Galinstan in a channel with an electrolyte such as HCL or NaOH as shown in Figure 2.6 [28]. Applying a bias down this channel causes the liquid metal to move towards the positive electrode. During biasing a skin develops on the side of the Galinstan slug close to the negative electrode, causing the Galinstan to get stuck. Removing the bias allows the electrolyte to dissolve the skin and a reverse bias speeds up the process. In a 2mm wide, $400\mu\text{m}$ high channel it is shown that a square wave signal with 8V peak-peak and a 3 VDC offset allows the metal to move without getting snagged by its own skin [30]. The same paper reports using circular chambers, shown in Figure 2.7, rather than a smooth channel to accurately control the position of the metal. This utilizes the high tension of the liquid metal to localize it at points where it can minimize its surface energy. It should be noted that the electrolytes used are usually lossy at microwave frequencies.

Figure 2.6: Continuous electrowetting with Galinstan in channel with electrolyte. Unbiased channel shown in (a) and biased channel shown in (b) causes liquid metal to move [28].

Figure 2.7: Liquid metal reservoirs. The Galinstan (Dark grey) moves from left to right [30].

CEW has been used to realize an amplifier with $\lambda/8$ double stub tuners. The double stubs are realized using a series of circular reservoirs to accurately control the position of the liquid metal. The amplifier can be tuned from 3.37-6.02GHz [35].

Electrochemically controlled capillarity (ECC) allows for variable length lines and this is not possible using the CEW method. The ECC method utilizes an electrolyte the same as the CEW method but utilizes the skin generated with electrical bias in an advantageous manner. Figure 2.8 shows how this method of moving the liquid metal functions. A positive voltage applied to the Galinstan reservoir and a negative voltage applied to the electrolyte results in an oxide forming. This oxide changes the surface tension to be close to zero and the liquid metal is pushed out of the reservoir and into the capillary. When a voltage is applied with the opposite polarity the oxide is removed and the high surface tension results in the Galinstan being pulled back into the reservoir. The liquid metal moves faster when being pulled into the reservoir

than being pushed out of it. This is likely because of the additional friction caused by the surface oxide [32].

Figure 2.8: Electrochemically controlled capillarity. The Galinstan will form an oxide skin and move out of the reservoir into the capillary with the voltage bias as shown in (a). Reversing the bias causes the skin to be dissolved and the Galinstan moves out of the capillary into the reservoir as shown in (b) [32].

The ECC method is used to realize a tunable monopole antenna that is demonstrated to function from 0.66-3.4GHz. It is shown that the antenna efficiency ranged from 41-70%, which is significantly lower than the > 90% efficiency typically seen for solid conductor antennas. This loss is due to the NaOH electrolyte solution which has a conductivity of 15S/m [32].

Memristor Switching

A memristor is by definition the fourth basic electronic component. It is a two terminal device with a resistance dependent upon the history of the applied voltage and/or current applied to the device. Recently, a microwave switch based on a memristor is presented [36]. The device is fabricated on 380nm silicon oxide grown on an intrinsic silicon substrate with two metal electrodes separated by a 35nm air gap. One electrode consists of three layers: the first is an adhesion layer of titanium; the second is 140nm thick gold; and the top layer is 70nm thick silver. The other electrode is 70nm thick silver. Applying a voltage ramp of 0-3V on the silver only electrode at the

higher potential causes a $110nm$ wide continuous silver filament to form from the silver side to the other electrode, thus turning on the switch. The switch is turned off with a ramp of 0 to $-0.4V$ which causes the filament to rupture. The magnitude of the voltage ramp determines the width of the filament and thus the resistance of the device. No microwave induced self-switching is observed in the off-state, however self switching is observed in the on-state with a power level above $17dBm$ regardless of frequency. The devices are reported to have an average $R_{ON} = 3.6\Omega$ and a $C_{OFF} = 1.4fF$, which corresponds to a figure of merit of $5fs$.

2.2.2 Semiconductor Switching and Tuning

Semiconductors are commonly used to realize components required for reconfigurable circuits. Semiconductors are advantageous to use because the entire circuit can easily be integrated. MEMS allow for the same advantage, but integration of MEMS usually requires many extra processing steps and can be problematic due to processing temperatures required to make the MEMS devices [9].

Semiconductor Switching

Switches are common semiconductor components and are frequently one of the first components designed in any semiconductor process. Typical CMOS switches are not advantageous for microwave applications and usually require a special process to function well at microwave frequencies. Silicon on insulator (SOI) and silicon on sapphire (SOS) are used to realize high performance switches. For these state-of-the-art switches it is possible to achieve an on-resistance of $0.3 - 0.5\Omega$, however it is unfeasible to achieve resistance as low as 0.1Ω in SOI. Switches are commonly characterized by a figure of merit which is $R_{ON} \times C_{OFF}$ and SOI switches have a value of $150 - 240fs$ [37]. Gallium arsenide based switches exhibit similar performance to

SOI based switches with a figure of merit of $270fs$ [12]. Silicon and gallium arsenide based devices have power handling problems which can be overcome with the use of gallium nitride based HEMT switches [38]. The switches have a figure of merit of $290fs$, which is comparable with other semiconductor switches [39]. PIN diodes can also be used when high power handling, good linearity, and lower loss than semiconductor switches are required [40–43]. PIN diodes are current biased devices and require more complex bias circuitry and power consumption than other semiconductor switches. The PIN diode switch has a figure of merit from $110 - 220fs$ [12].

Recently, a new semiconductor switch is presented and is being referred to as the Super-Lattice Castellated Field Effect Transistor (SLCFET) [44–46]. The development of this switch involved creating a custom semiconductor process for GaN and the HEMT device structure as shown in Figure 2.9 [44]. The structure consists of stacked layers of the 2D electron gas (2DEG) formed at the interface of AlGaN/GaN junctions. In order to be able to fully deplete the 2DEG, the gate is wrapped around the channel and multiple parallel stacked channels are used to lower the on-state resistance. The device and process are optimized to realize high performance microwave switches with an $R_{ON} = 0.41\Omega/mm$ and $C_{OFF} = 0.19pF/mm$. This gives a figure of merit of $78fs$, which is better than SOI switches and the device exhibits power handling capability of typical GaN devices.

Figure 2.9: Basic structure of SLCFET [44].

Varactor Tuning

Semiconductor based varactor diodes are used extensively to provide a method to tune the response of circuits [20, 47–51]. These components have a similar loss to that of PIN diode switches when no bias is applied. The loss is based on the reverse bias on the varactor. Under a large reverse bias the series resistance of the varactor becomes a fraction of an Ohm. Tuning ratios in the range of 25:1 can be seen for semiconductor varactors [52].

Optical Tuning

Another method of using semiconductor devices is in conjunction with an optical control signal. The optical signal generates charge carriers in the semiconductor which can be used to generate a voltage to tune a varactor [53] or activate a switch [54]. Alternatively, the generated charge carriers can be used directly to modify the conductivity [55, 56] of the material or modify the effective permittivity [57, 58]. The modification of the effective permittivity is different from the method used at optical frequencies. This method involves semiconductor particles with diameters in the

micrometer range dispersed in an insulating material. Applying light to the material generates charge carriers, which causes the semiconductor to change from a dielectric to a conductor. These localized areas of conductivity result in realizing an artificial dielectric very similar to the metal sphere loaded materials investigated more than half a century ago [59]. This method is limited to a maximum of $1.4\times$ increase in effective index of refraction [59].

2.2.3 Tunable Materials

Materials with tunable or switchable material properties provide another excellent path to realizing reconfigurable microwave circuits. The most common materials found in literature are phase change materials, ferrites, ferroelectrics, liquid nematic crystal, and metamaterial. Metamaterial will be omitted in this section and discussed in depth in the next section.

Phase Change Materials

Phase change materials are classified as materials that can be reversibly changed from one phase to another, usually between insulating and conductive phases, which can correspond with many orders of magnitude change in conductivity. Phase transitions have been demonstrated using optical, electrical, stress, and thermal control signals. Vanadium dioxide is one phase change material which has a phase transition that can be controlled using the four methods mentioned [60–63]. A recently demonstrated switch using VO_2 achieved a figure of merit of less than $15fs$. The switch is realized using a CPW line with a $3 \times 20\mu m$ gap with $200nm$ thick VO_2 deposited in the gap. The switch has an R_{ON} of 2Ω and a C_{OFF} of $7.7fF$. The paper shows that the switch can also be electrically actuated with a current of $20mA$ applied through the two switch terminals [64].

Germanium telluride is another phase change material (PCM) that can be thermally changed between an amorphous and crystalline state which corresponds with a change from an insulator to a conductor. Switching between amorphous and crystalline states requires different temperature profiles to be applied to the material, which is usually accomplished with a microheater embedded close to the phase change material. Changing the material to the amorphous phase (insulating) requires a quick increase in temperature to 725°C , causing the PCM to melt, then quickly quenching the material. Changing the material to the crystalline phase (conducting) requires a temperature above 190°C but lower than the melting temperature. This temperature is applied for a longer time than when switching to the amorphous state, and is ramped down slowly. For one switch reported, switching between on or off requires applying 0.5W or 4W for $1.5\mu\text{s}$ or 30ns and the switch shows an RF power handling capability of 3.1W [65]. One fabricated switch has an $R_{ON} = 0.9\Omega$ and $C_{OFF} = 14.1\text{fF}$, which gives a figure of merit of 13fs [66]. This switch is implemented as shown in Figure 2.10 [65]. The PCM used is $0.9\mu\text{m}$ long and $30\mu\text{m}$ wide. PCM switches for microwave applications have progressed significantly over recent years and have demonstrated excellent characteristics for use in communication systems [65–75].

Figure 2.10: Cross section of phase change switch shown in (a) and top view shown in (b) [65].

Ferrites

Ferrite materials have been used for many years in applications that require phase shifting of signals. These materials have a magnetically controllable permeability. The main problems with this class of material are: the physical size of the setup; the large magnets or electromagnets that are required; and the cost. There are some reports of miniaturized ferrite devices with the use of LTCC ferrite tapes [41, 76, 77].

Ferroelectrics

Ferroelectric materials are the dual of ferrite materials; the permittivity of the material is controlled with an electric bias field. These materials suffer from needing extremely high intensity field, in the range of MV/m, and are temperature sensitive [78, 79].

Liquid Nematic Crystal

Liquid nematic crystals which are commonly used in liquid crystal displays also provide unique capabilities [80–82]. The electrical properties of these materials are normally changed by applying an electrical signal that causes molecules in the material to change alignment and in turn allow control of the dielectric constant of the material.

2.2.4 Summary

There are numerous methods shown to achieve tunable and reconfigurable microwave circuits. The most promising methods include MEMS switches and varactors, Galinstan based microfluidics, memristors, semiconductor switches and varactors, and GeTe based PCM switches. The development of new microwave switches

has been extensive in recent years and is one of the key developments required to practically implement highly reconfigurable or programmable microwave circuits.

2.3 Metamaterials

Metamaterials are a class of engineered materials that usually utilize metallic structuring to design a material with specific effective permittivity and permeability. This structuring is usually smaller than a tenth of the wavelength of the highest frequency electromagnetic wave of interest. This structuring allows for almost any value of permittivity or permeability to be engineered. A wide variety of metamaterials are reported in the literature [83,84]. Metamaterials allow for the realization of electromagnetic invisibility cloaks and lenses that can image beyond the diffraction limit [85]. In addition, metamaterials are good for miniaturizing some circuits such as resonators by taking advantage of double negative metamaterials having opposite phase progression as that of double positive materials [86]. These materials can be broken into four categories as shown in Figure 2.11 [84]. When analyzing metamaterials, all parts of the unit cell (i.e. metallic structures, active components, and substrate) are represented as an effective permittivity and permeability, which significantly simplifies circuit design using these structures. Most of the literature has focused on designing double negative metamaterials. These are materials which have simultaneously negative permittivity and negative permeability over a small frequency range. Researchers focus on this because these materials provide new functionality not possible with naturally occurring materials. Artificial materials with both positive permittivity and positive permeability receive less attention, but are potentially useful as well.

Figure 2.11: Permittivity-Permeability Diagram [84].

Metamaterials can be analyzed using a transmission line approach [83], which yields the ideal unit cells shown in Figure 2.12 for each type of material [87].

Figure 2.12: Equivalent circuits for ideal transmission lines with: (a) $\mu > 0$ and $\epsilon > 0$; (b) $\mu > 0$ and $\epsilon < 0$; (c) $\mu < 0$ and $\epsilon > 0$; (d) $\mu < 0$ and $\epsilon < 0$. Δz represents a length of transmission line much less than λ [87].

One practical implementation of a negative permittivity metamaterial is an array of metal wires parallel to the orientation of the electric field of the incident EM field [88]. One of the most common types of structures used to create a negative permeability is known as the split-ring resonator. This structure consists of two circular rings, one inside the other, with each broken in one spot to form a 'C'; the inner one is broken on the opposite side as the outer one [84]. This structure only exhibits a negative permeability in a narrow frequency range just after self-resonance of the structure. Prior to self resonance the split ring causes the effective permeability of the material to increase rapidly versus frequency. The split ring resonator combined with the wire media can make a double negative metamaterial as shown in Figure 2.13 [89].

Figure 2.13: Unit cell of first double negative metamaterial presented. Both structures shown are metal. Stacking of the cells would connect the metal posts for a continuous post, while the split ring resonators are electrically not connected.

Reconfigurable and tunable metamaterial are presented extensively in the literature [31, 61, 62, 90–99]. There are reconfigurable metamaterials presented using almost every tuning/reconfiguration method presented previously [100]. However, demonstrations of fully programmable substrates are currently not reported in the literature. There is one programmable metasurface presented with the ability to be programmed with relatively fine resolution to manipulate EM waves [101]. Double negative metamaterials are of less interest to this work, and for this reason an extensive discussion of the reconfigurable and tunable double negative metamaterial is not given.

Double positive metamaterials have received relatively little attention, most likely because materials with a wide range of dielectric and magnetic properties can be made by mixing together naturally occurring materials. Creating metamaterials with an artificially higher permittivity is seen more often in literature than creating materials with artificially higher permeabilities. The most common structure used has an 'I' shape or a dumbbell shape as shown in Figure 2.14 [102–110].

Figure 2.14: Structure used to make material with higher effective permittivity.

Figure 2.15: Substrate composed of dumbbell structures.

A vertical and horizontal array of these dumbbell structures with an incident electric field oriented as shown in Figure 2.14 will appear as though it has a significantly higher permittivity. A material composed of dumbbells would look similar to that shown in Figure 2.15. The dumbbell structure is meant to function similarly to the spherical metallic particles used to realize artificial dielectrics. Circulating currents

in the spherical metallic particles caused a decrease in effective magnetic constant with an increase in effective dielectric constant [102]. The resulting refractive index could only be increased by 40% [59]. The dumbbell particle significantly reduces circulating current which allows for large increases in dielectric constant with minimal reduction in magnetic constant [102, 103].

Reconfigurable and tunable artificial dielectrics are not easily located in the literature and frequently take the form of a tunable or reconfigurable transmission line. One paper presents a differential transmission line that is designed with a digitally controlled phase shifting capability and is shown in Figure 2.16 [111, 112]. The effective dielectric constant that the transmission line sees can be increased, up to twice the original value, by shorting together the metal strips that are perpendicular to the transmission line using a switch.

Figure 2.16: Digitally tunable differential transmission line [112].

Transmission lines can be represented using an equivalent circuit unit cell. If these unit cells represent an electrical section much less than λ in length then they reproduce the properties of the transmission line accurately. Circuits can be constructed

using these unit cells with lumped components rather than the transmission lines. This allows for microwave structures such as rat-race rings, quadrature directional couplers, and phase shifters to be constructed using lumped elements which significantly decreases circuit size. In addition to decreased size, components such as varactors can be substituted for capacitors to make the circuits tunable, or switches can be added to make the circuits reconfigurable [49].

2.4 Reconfigurable Circuits

There has been a plethora of reconfigurable microwave circuits demonstrated for the majority of the microwave devices required for a microwave system. However, the majority are realizing a single reconfigurable component and are not of significant interest for this work. Matching circuits are useful in many application and are reviewed. Next, circuits that can be considered field programmable microwave circuits are shown, which include programmable waveguides and transceiver chains. Since this area is new many of the examples given have not been demonstrated in the literature and only a patent exists for the idea. Next, a quick review of some commercially developed programmable radios is given. Finally, substrate integrated waveguide circuits are very similar to the waveguides presented in this work and are thus reviewed.

2.4.1 Programmable Matching Circuits

A variety of matching circuits are presented in the literature that can match an input impedance almost anywhere on a Smith chart [21, 113–115]. One tuner uses twelve open-circuited stubs that can be connected to a transmission line using MEMS switches as shown in Figure 2.17. This tuner is designed to match any reflection

coefficient between 0-0.5, and can match some reflection coefficients greater than 0.5. The match is better than 10dB.

Figure 2.17: Brute force 12-bit tuner using open-circuited $l=75$ mil stubs which are spaced $d=440$ mil apart [114].

These reconfigurable matching circuits can be used to realize reconfigurable amplifiers [18,20], or used anywhere a reconfigurable matching circuits are advantageous.

2.4.2 Field Programmable Microwave Circuit

The earliest field programmable microwave circuit (FPMC) found in literature is reported by Bozler et al. and is a MEMS switch array to realize programmable distributed components [116]. The circuit consists of an array of square metal patches with a pitch of $100\mu\text{m}$ that can be connected together using MEMS switches as shown in Figure 2.18

Figure 2.18: Envisioned layout for reconfigurable distributed components. Metal patches are at least twenty times smaller than the wavelength of interest and the patches can be connected dynamically using MEMS switches [116].

An electrically functional mock-up of this concept is created using different metal patterning to represent open- and closed MEMS switches. These electrically similar structures to the MEMS switches are then used in conjunction with metal patches to demonstrate matching networks, power amplifiers, and filters [116]. Further demonstration of fully reconfigurable distributed components is not reported in the literature. Similar to this work is a patent that uses thermal phase change material rather than MEMS to dynamically wire together the metal patches [117]. Along the same line, there are pixel antennas presented which show the ability to reconfigure from $1 - 6GHz$ [118]. Due to the switches being in a series configuration these antennas have much higher losses than their non-reconfigurable counterparts. Rockwell

Collins has a patent for a similar pixel antenna that uses photoconductive switches to dynamically connect together the metal patches [119].

Another FPMC proposed is called the programmable microwave function array (PROMFA). A single PROMFA cell has four bi-directional inputs that can be switched on or off. Each cell consists of four sub cells as shown in Figure 2.19. Each of these sub cells consists of a switch, and two directional amplifiers. Each sub cell amplifier has an array of transistors that can be individually activated. This allows for similar behaviour as changing the transistor size, which allows for flexibility in the frequency domain. Using many of these PROMFA cells the authors demonstrate power splitter/combiners, routers, amplifiers, phase shifters, time delay lines, corporate feed networks, and filters over 4-8 GHz. No noise or linearity analysis is shown [120, 121].

Figure 2.19: Block diagram for PROMFA on left and sub cell with transistor array on right [120, 121].

A different group present another block that is multi-functional and to be used in the PROMFA. The device has three ports and consists of a single resonant metal patch. Depending on how the patch is excited it acts as either a patch antenna or a filter and is shown in Figure 2.20 [122].

Figure 2.20: In (a) filter performance is between port 1 and 2 (front), while in (b) antenna performance is achieved by exciting port 3 (back) [122].

Lockheed Martin recently filed a patent for a field programmable microwave array that consists of a switch matrix to dynamically wire together microwave circuits from a selection of pre-designed microwave circuits [123]. Another patent recently filed by Raytheon is for optically programmable transmission lines. This involves a photoconductive layer that could be selectively illuminated to change areas from insulating to conducting so as to dynamically realize transmission lines [124].

DARPA has a project entitled adaptive RF technologies (ART). The goal of this project is to realize field programmable communications devices up to $30 + GHz$ that are indifferent to specific waveforms and standards. A main thrust under the ART project is the RF-FPGA. This project aims to capture the performance of multiple microwave transceivers within one device. An expected realization involves a transceiver chain that is programmed by dynamically rewiring connections between basic microwave circuit blocks that have a certain level of reconfigurability [125]. Under the RF-FPGA thrust Northrop Grumman received funding for a $0.4 - 18GHz$ multi-functional transceiver. The proposed device consists of banks of components that are dynamically connected together [126]. The proposal involves the development of multi-pole multi-throw PCM switches which are under active

development [66]. Under this same DARPA project BAE Systems received funding and has recently reported a programmable microwave front end that functions from 0.4-20GHz [127]. This system uses a plurality of reconfigurable microwave components in a flexible switch matrix. The design includes multi-functional 1 – 20GHz blocks that can act as amplifiers, frequency converters, and 4-way switches for signal routing. The design also consists of $DC - 6GHz$ blocks which perform amplification, filtering, up/down conversion, and signal routing. There are also 0.01 – 20GHz configurable frequency generator blocks. The noise figure after the first LNA is given to be 10 – 12.5dB [127]. This approach from BAE Systems appears to be similar to the PROMFA shown earlier.

2.4.3 Commercial Programmable Radios

Some commercial radios have been demonstrated that can either cover extremely large bandwidths or are programmable. Lime Microsystems has a commercially available programmable radio frequency integrated circuit. This circuit can function from 100kHz – 3.8GHz and they have a version planned to operate from 100kHz – 12GHz [4]. Analog Devices also produces a frequency agile transceiver that functions from 70MHz to 6GHz [5]. Recently, there is an all digital transmitter demonstrated that is able to create fourteen different cellular base station signals without the use of a digital to analog converter. The group did state realizing a receiver would be an order of magnitude more difficult [128].

2.5 Substrate Integrated Waveguide Circuits

The substrate integrated waveguide (SIW) is a waveguide that consists of solid metal top and bottom conductors and metal posts/vias for side walls as shown in Figure

2.21 [129]. The SIW is normally interpreted as being equivalent to the rectangular waveguide [130]. The equivalence requires the calculation of an effective width (w_{eff}) given by Equation 2.1 .

Figure 2.21: A simple substrate integrated waveguide.

$$w = \frac{2w_{eff}}{\pi} \cot^{-1} \left(\frac{\pi d}{4w_{eff}} \ln \frac{s}{d} \right) \quad (2.1)$$

where,

d is the diameter of the vias/posts

s is the pitch of the posts

w is the width of the waveguide (center of shorted post to center of shorted post)

There are various other equations reported in literature for the effective width of the waveguide [129]. One paper presents an interpretation of the SIW as a slab waveguide. It is shown that the post walls of the waveguide resembles that of a negative permittivity material. Using the negative permittivity material values in the slab waveguide equations results in the same propagation characteristics as measured and simulated waveguides [131]. The circuits utilized in this thesis have

the configuration of a slab waveguide with a positive permittivity core sandwiched between negative permittivity material. For this reason reconfigurable SIW circuits are reviewed.

In a paper that reviews tunable SIW resonators, six basic tuning methods are given [132]. These include reactively loading the sidewall of a gap in the cavity, a switchable post in the cavity, a capacitively loaded post cavity, a diode loaded floating patch, ferrite loaded cavity, and mechanical tuning (i.e. screw flap). The SIW circuits of interest for this thesis are the post loaded or switched circuits.

One reported SIW filter is made reconfigurable by placing a post within the SIW cavity at a location of high electric field. Rather than shorting the two plates together, the post is passed through the top metal and a switch is connected between the top metal plate and the post as shown in Figure 2.22. The filter is tuned from $1.55\text{--}2\text{GHz}$ with an unloaded quality factor ranging from $120 - 101$. The filter is fabricated on RT/Duroid 6010. A significant amount of the loss is attributed to the PIN diodes [133]. A slightly different filter is fabricated by the same group, but off-the-shelf MEMS switches from Omron Corporation are used rather than PIN diodes. The filter uses a substrate with identical properties for the cavities with the switches in slightly different locations. The filter is tuned from $1.2\text{--}1.6\text{GHz}$ and has an unloaded quality factor from $132 - 93$. The characteristics of the filter are very similar to the PIN diode filter likely because the MEMS switches have an $R_{ON} = 1.5\Omega$ which is similar to that of PIN diodes [134].

Figure 2.22: Tunable SIW filter. (a) is top view and (b) is side view of PIN diode loaded post [133].

Using a very similar cavity loading structure to that shown in Figure 2.22b and loading it with a varactor rather than a switch allows for a VCO to be realized. The VCO is tuned from 1.7-2.2GHz and had a phase noise better than $-109dBc/Hz$ at 100kHz offset over the tuning range [135].

A reconfigurable resonant cavity is designed with mechanically switchable posts. Open- or short circuiting the posts as shown in Figure 2.23 allows for the resonant frequency of the cavity to be changed. The resonant cavity is fabricated using 31 mil Rogers RT/Duroid 5880. In the on-state the cavity resonates at 10.6GHz with a Q of 448, while in the off-state the cavity resonates at 8.63GHz with a Q of 274 [136].

Figure 2.23: Reconfigurable SIW resonant cavity utilizing switched posts. (a) On-state; (b) Off-State [136].

2.6 Conclusions

A variety of methods to reconfigure and tune microwave circuits are presented. Many of these methods show extremely promising characteristics and it is likely that MEMS switches and varactors, Galinstan based microfluidics, memristors, semiconductor switches and varactors, and GeTe based PCM switches will see use in high performance FPMCs. Metamaterials are reviewed and shown to provide an avenue to realizing reconfigurable or tunable microwave circuits, potentially in the form of a programmable substrate. Finally, reconfigurable circuits are presented. These included: the relatively mature impedance matching circuits; the nascent field of FPMC; commercial programmable transceivers; and substrate integrated waveguides. The review of FPMCs includes transceivers with rewirable block chains and programmable cells with basic microwave circuits such as amplifiers. There are programmable waveguides presented, however they are either just ideas or limited in functionality. In the following chapter it will be illustrated that the configuration of these waveguides will lead to undesirably high losses. The commercial transceivers

demonstrate a current demand for programmable radios, but they are limited in upper frequency of operation and programmability. The SIW circuits are presented as being similar in function to the programmable waveguides that will be detailed in this work. Some tunable SIW circuits presented have promising characteristics.

Chapter 3

Programmable Waveguides

3.1 Introduction

A significant amount of microwave circuit design revolves around the use of wave guiding structures. This makes the waveguide one of the most general purpose microwave components and is the first component required to realize programmable microwave circuits. This dissertation presents a method to implement programmable waveguides by realizing a field programmable microwave substrate. The substrate is made up of square unit cells that are significantly smaller than the wavelength of the highest frequency of operation. The effective dielectric constant of these unit cells can be switched between positive and negative dielectric states. It is also desirable for the unit cell to synthesize a variety of positive dielectric constants.

It is well known that a transmission line can be represented by an equivalent circuit as shown in Figure 3.1. This equivalent circuit can be used to realize an artificial transmission line. The circuit shown is the 1D equivalent circuit for a transmission line. For implementation of an artificial transmission line it may be more useful to use a 2D equivalent circuit. The 2D equivalent circuit is almost the same as that shown in Figure 3.1, but additional series inductors and resistors of the same values

are used 90° to those shown. Depending on the type of transmission line the values for R , L , G , and C are different.

Figure 3.1: Equivalent circuit of transmission line for a small length Δz .

For the purposes of analysis in this chapter the equivalent circuit equations for a parallel plate transmission line are used, and are given in Equations 3.1-3.5 [137].

$$L = \frac{\mu d}{W} \quad (H/m) \quad (3.1)$$

$$C = \frac{\epsilon' W}{d} \quad (F/m) \quad (3.2)$$

$$R = \frac{2R_s}{W} \quad (\Omega/m) \quad (3.3)$$

$$G = \frac{\omega \epsilon'' W}{d} \quad (S/m) \quad (3.4)$$

$$R_s = \sqrt{\frac{\omega \mu}{2\sigma}} \quad (\Omega/m) \quad (3.5)$$

where,

- μ is the permeability of the material filling the transmission line.
- $\epsilon = \epsilon' - j\epsilon''$ is the permittivity of the material filling the transmission line.
- W is the width of the transmission line or a section of the transmission line.
- d is the distance between the two parallel plates.
- σ is the conductivity of the metal parallel plates.
- ω is the frequency of operation.

For the equivalent circuit to function as a transmission line, Δz must correspond to $\lambda/8$ or less at the highest desired frequency of operation. A Δz larger than $\lambda/8$ can be used, but the behavior deviates significantly from that of the intended transmission line. For the purposes of this work W is the same as Δz (square). The Nicolson-Ross-Weir (NRW) algorithm given in Equations 3.6-3.10 [138] is used to calculate the effective permittivity and permeability using s-parameters of the transmission line equivalent circuit.

$$\Gamma = \frac{S_{11}^2 - S_{21}^2 + 1}{2S_{11}} \pm \sqrt{\left(\frac{S_{11}^2 - S_{21}^2 + 1}{2S_{11}}\right)^2 - 1} \quad (3.6)$$

$$T = \frac{S_{11} + S_{21} - \Gamma}{1 - (S_{11} + S_{21})\Gamma} \quad (3.7)$$

$$\frac{1}{\Lambda^2} = - \left[\frac{1}{2\pi\ell} \left(\ln \left(\frac{1}{T} \right) + i2\pi n \right) \right]^2 \quad (3.8)$$

$$\mu_r = \frac{1 + \Gamma}{(1 - \Gamma)\Lambda \sqrt{1/\lambda_0^2 - 1/\lambda_c^2}} \quad (3.9)$$

$$\epsilon_r = \frac{\lambda_0^2}{\mu_r [1/\lambda_c^2 - 1/\Lambda^2]} \quad (3.10)$$

Where:

- Γ is the reflection coefficient and the correct choice of sign in Equation 3.6 is made by ensuring $|\Gamma| \leq 1$.
- T is the transmission coefficient.
- μ_r is the relative permeability.
- ϵ_r is the relative permittivity.
- S_{11} and S_{21} are the reflection and transmission s-parameters.
- λ_0 is the free-space wavelength.
- λ_c is the cutoff wavelength of the transmission line.
- ℓ is the length of the sample and should be less than 180 electrical degrees.
- $n=0, \pm 1, \pm 2, \dots$
- The term $i2\pi n$ in Equation 3.8 is used to unwrap the imaginary part of $\ln(\frac{1}{T})$.

This chapter analyzes positioning of a lossy component in a transmission line equivalent circuit (unit cell). The lossy component represents a switch that is utilized to realize programmable behavior of the unit cell. The analysis aims to identify optimal placement and methods to mitigate losses from the component. This analysis is performed for both positive and negative dielectric materials. The next section discusses propagation characteristics of slab waveguides using the materials described in the previous section. The conclusions drawn in this chapter are used to design the unit cell for the FPMS in later chapters.

3.2 Analysis of Unit Cells

The work presented in this thesis requires a material with programmable permittivity or permeability. A programmable material is realized by using a 2D array of unit cells that can be represented as transmission line equivalent circuits. Dynamically changing the components in the equivalent circuit allows for a programmable material to be implemented. These components are switched in and out with the use of switches, however the switch introduces unwanted losses. This requires the identification of a switch position that allows for methods to minimize and mitigate unwanted losses. To perform this analysis, a 5Ω resistor is used to represent a switch in the on-state. This resistance is larger than the best reported in the literature, but the larger value is used to better illustrate losses and demonstrate decent performance can still be obtained with low performance switches. The analysis is simplified by assuming a lossless material, which allows for the losses from the switch to be easily identified and minimized. The off-state characteristics of the switch are significantly less of an issue and are not included in this analysis. The transmission line equivalent circuits used for analysis are shown in Figure 3.2. Unless stated otherwise the values of the inductor and capacitor are calculated using Equations 3.1 and 3.2 and the material used for the calculations is a lossless $1mm$ thick substrate with a dielectric constant of 4.9.

Figure 3.2: Schematics for simulating effect of a lossy component on transmission line unit cell. (a) is the switch in the series configuration and (b) is the switch in the shunt configuration.

The terminal impedance in Figure 3.2 is calculated using Equation 3.11.

$$Z_{Term} = \sqrt{\frac{L_o}{C_o}} \quad (3.11)$$

Where: L_o and C_o are calculated using Equations 3.1 and 3.2 with ϵ and μ the values for vacuum permittivity and permeability.

3.2.1 Positive Dielectric

The unit cells given in Figure 3.2 are simulated using Keysight Advanced Design System (ADS) and the effective properties are extracted using the NRW algorithm. This first simulation performed is for unit cells with sizes of 1, 2 or 5 mm and the properties are shown in Figures 3.3 and 3.4. The capacitance, inductance, and terminal impedance for 1, 2, or 5mm unit cells are given in Table 3.1. It should be noted that all dielectric and magnetic losses are added together when presented. This allows one to more easily interpret the results. For some of the results the dielectric loss value increases with frequency and the magnetic loss value decreases below zero. A negative magnetic loss value has been observed in metamaterials and is considered

to be an accurate description [139].

Unit Cell Size (mm)	C (fF)	L (nH)	Zterm (Ohm)
1	43.39	1.26	376.73
2	173.54	1.26	188.37
5	1084.64	1.26	75.35

Table 3.1: Values used for simulating transmission line equivalent circuit.

Figure 3.3: Material properties extracted from s-parameter simulation of unit cell with lossy component in the shunt or series configuration.

Figure 3.4: Loss tangents extracted from s-parameter simulation of unit cells with the lossy component in the shunt or series configuration.

The results shown in Figures 3.3a and 3.3b demonstrate that a smaller unit cell better approximates the transmission line it is meant to represent. In this case it is supposed to represent a parallel plate transmission line filled with a substrate that has a dielectric constant of 4.9 and magnetic constant of 1. In Figure 3.4 it can be seen that the switch in the series configuration results in significant losses, and reducing the cell size has no affect on losses. However, the switch in the shunt configuration results in a significant reduction in loss when the unit cell is shrunk. The switch in the series configuration is the same as that demonstrated by a few authors for programmable waveguides [116, 117, 124] or pixel antennas [118, 119]. This simple comparison shows that unless the switches have extremely small loss, the overall loss of the system will be too high. The programmable waveguide (i.e. small metal patches that are dynamically connected together using MEMS switches) presented by Bozler et al. [116] does not analyze their structure using effective material properties. Analysis of this structure using effective material properties shows that the switch changes the unit cell from a double positive material when in the on-state to a negative permeability material when the switch is in the off-state.

The effect of different dielectrics are analyzed next. This is the same as modifying the shunt capacitance in the transmission line equivalent circuit. A simulation is performed using a dielectric constant of 2 and the results are shown in Figures 3.5 and 3.6.

Figure 3.5: Material properties extracted from s-parameter simulation of unit cell using a substrate with $\epsilon_r = 2$.

Figure 3.6: Loss tangents extracted from s-parameter simulation of unit cell using a substrate with $\epsilon_r = 2$.

The results in Figure 3.5 show that effective material properties better approximate the intended properties than the previously analyzed cell. This is because the cell has a lower dielectric constant and thus is electrically smaller than the previous cell that used a dielectric constant of 4.9. As mentioned earlier, a smaller cell size electrically allows for a better approximation of the parallel plate transmission line. Comparing Figure 3.6 and 3.4 it can be seen that the loss with the switch in the series configuration is the same. However, the loss is decreased for the switch in the shunt

configuration. This implies that the loss contribution from the switch is affected by the shunt capacitance. To analyze this further, the next simulations are performed for the cell with a dielectric constant of 10 and the results are shown in Figures 3.7 and 3.8.

Figure 3.7: Material properties extracted from s-parameter simulation of unit cell using a substrate with $\epsilon_r = 10$.

Figure 3.8: Loss tangents extracted from s-parameter simulation of unit cell using a substrate with $\epsilon_r = 10$.

For the 5mm cell the effective permittivity and permeability shown in Figure 3.7 shows a large deviation from the intended properties as the frequency increases.

This is because $\lambda_g/8 = 2.34mm$ at 5GHz, which means the equivalent circuit approximation for the 5mm unit cell becomes very inaccurate. The properties for the 1mm and 2mm unit cells match much more closely with the intended values ($\epsilon = 10$ and $\mu = 1$). The loss when the switch is in the series configuration is not affected by the shunt capacitance as seen in Figures 3.4, 3.6, and 3.8. For the switch in the shunt configuration it can be seen that the losses are greater for the transmission line with a dielectric constant of 10 than the lines with a dielectric constant of 2 and 4.9. These analyses show that decreasing the size of the shunt capacitor results in the switch contributing less to the loss of the unit cell.

Further investigation of the effect of the shunt capacitor (dielectric constant) is performed at a frequency of 1GHz. The following simulation utilized a 1mm unit cell and sweeps the shunt capacitance from 0.01 – 10pF. As seen in Figure 3.9a, the dielectric constant scales linearly with the increase in capacitance. Figure 3.9b shows the capacitance has no affect on the magnetic constant. It can be seen in Figure 3.10 that the losses are heavily dependent on the shunt capacitance value and become prohibitively large with capacitance values larger than 2pF.

Figure 3.9: Material properties extracted from s-parameter simulation of unit cell for various shunt capacitance values.

Figure 3.10: Loss tangents extracted from s-parameter simulation of unit cell for various shunt capacitance values.

More analysis of the shunt capacitor is performed with a simulation in which the shunt capacitance is 0.1pF and the unit cell size is changed. The change in unit cell size is done by varying the impedance of Z_{Term} . Since inductance only varies with distance between the parallel plates this results in behavior equivalent to changing the effective dielectric constant of the material as can be seen in Figure 3.11a. The magnetic constant shown in Figure 3.11b is unity as expected. The results shown in Figure 3.12 show that the unit cell size has no affect on the loss contribution from the shunt resistor.

Figure 3.11: Material properties extracted from s-parameter simulation of unit cell with shunt capacitance of $0.1pF$, inductance of $1.257nH$, and varying unit cell size.

Figure 3.12: Loss tangents extracted from s-parameter simulation of unit cell with shunt capacitance of $0.1pF$, inductance of $1.257nH$, and varying unit cell size.

The results given so far have mainly been investigating the effect of the shunt capacitor on losses from the resistor used to represent the switch. These results show that the unit cell size does not affect the loss contributions from the switch. The only parameter that reduces the loss contribution from the switch has been the shunt capacitor. The characteristic impedance of a transmission line is calculated as $Z_o = \sqrt{L/C}$. The decrease in capacitance results in a larger characteristic impedance for the transmission line. If there is a relation between loss contribution from the

switch and characteristic impedance of the transmission line, then increasing series inductance should decrease loss. It was seen that this is not the case. A simulation performed with a $1mm$ unit cell, shunt capacitance of $43.39fF$, shunt 5Ω resistor, and the series inductance swept from $0.01 - 10nH$ showed no change in either loss or dielectric constant. Only the magnetic constant increased with an increase in the series inductance.

The switch in the shunt configuration as shown in Figure 3.2b is not the same as the typical configuration for the transmission line equivalent circuit as shown in Figure 3.1. Converting the shunt capacitor (C_1) and resistor (R_1) in series to a shunt capacitor (C_2) and resistor (R_2) in parallel yields the values for capacitance and resistance given in Equations 3.12 and 3.13. From the equation for R_2 it can be seen that decreasing the capacitance C_1 results in a large R_2 . A larger value for the shunt resistance is the desired behavior as seen for low loss material using Equation 3.4. Using Equations 3.2, 3.4, 3.12, and 3.13 and solving for dielectric constant yields Equation 3.15 and the dielectric loss tangent yields Equation 3.14. Using this equation one is able to determine given a certain switch resistance, what capacitance is required to achieve a specific dielectric loss tangent.

$$C_2 = \frac{C_1}{C_1^2 R_1^2 \omega^2 + 1} \quad (3.12)$$

$$R_2 = \frac{C_1^2 R_1^2 \omega^2 + 1}{\omega^2 C_1^2 R_1} \quad (3.13)$$

$$\tan \delta = R_1 \omega C_1 \quad (3.14)$$

$$\epsilon' = \frac{C_1 d}{(C_1^2 R_1^2 \omega^2 + 1) W^2} \quad (3.15)$$

where,

- C_1 is the shunt capacitance of the transmission line equivalent circuit.
- R_1 is the resistor in series with the shunt capacitance.
- d is the parallel plate separation.
- W is the width of the parallel plate unit cell.

The lossy component should be placed in the shunt configuration, but there are still losses for the series inductance in a real circuit that should be minimized. The analysis is performed with a $1mm$ unit cell and the series inductance is swept from $0.01 - 10nH$. The resistor is in the series configuration for this simulation and the shunt capacitance remains constant. As seen in Figure 3.13a there is no change in dielectric constant, while the magnetic constant increases linearly with series inductance increase as shown in Figure 3.13b. It can be seen that a larger series inductance decreases the loss of a series resistance as shown in Figure 3.14.

Figure 3.13: Material properties extracted from s-parameter simulation of unit cell with lossy component in series configuration and sweeping series inductance value.

Figure 3.14: Loss tangents extracted from s-parameter simulation of unit cell with lossy component in series configuration and sweeping series inductance value.

Simulations performed with the lossy component in the series configuration and varying the unit cell size while maintaining the same series inductance and shunt capacitance resulted in no change in losses. A different analysis is performed for the lossy component in either the series or shunt configuration and the distance between parallel plates is swept. For one sweep, a cell of 1mm loaded with a substrate that has a dielectric constant of 4.9 is used and the plate separation distance is swept

from $0.1 - 3\text{mm}$. From the results seen in Figures 3.15 and 3.17 it can be seen that the properties remain constant. The loss for both configurations as seen in Figure 3.16 and 3.18 show a decrease in loss with an increase in separation. The increase in separation results in a decrease in capacitance and increase in inductance which was previously shown to reduce losses. The decrease in loss is exponential for both configurations of the lossy component which means that after a certain separation distance there is not much advantage in further increase.

Figure 3.15: Material properties extracted from s-parameter simulation of unit cell with lossy component in shunt configuration, sweeping plate separation, and maintaining $\mu_r = 1$ and $\epsilon_r = 4.9$.

Figure 3.16: Effective combined loss tangents extracted from s-parameter simulation of unit cell with lossy component in shunt configuration, sweeping plate separation, and maintaining $\mu_r = 1$ and $\epsilon_r = 4.9$.

Figure 3.17: Material properties extracted from s-parameter simulation of unit cell with lossy component in series configuration, sweeping plate separation, and maintaining $\mu_r = 1$ and $\epsilon_r = 4.9$.

Figure 3.18: Loss tangents extracted from s-parameter simulation of unit cell with lossy component in series configuration, sweeping plate separation, and maintaining $\mu_r = 1$ and $\epsilon_r = 4.9$.

3.2.2 Negative Dielectric

The positive dielectric material is used for the core of a waveguide while a negative dielectric material is used to confine the electromagnetic energy. The negative dielectric is realized by replacing the capacitor shown in Figure 3.2b with an inductor. As

in the previous section, the main goal is to minimize the loss from the shunt resistor that represents a switch.

The first analysis uses a $5mm$ unit cell with a shunt inductor with values ranging from $0.1 - 2nH$. The results shown in Figure 3.19a demonstrate that a larger inductance results in a more negative dielectric initially, but the situation at higher frequencies is a larger inductance results in a less negative value. The results shown in 3.19b show that the shunt inductor affects both the magnetic constant and the dielectric constant. This is different than the double positive materials, in which the shunt capacitor only affects the effective dielectric properties and the series inductor only affects the magnetic properties. The $0.1nH$ shunt inductor appears to causes different behavior versus frequency than the other inductor values as seen in Figure 3.19b. However, simulations up to $15GHz$ reveal the shape of the curves are similar. A larger shunt inductance results in lower losses as shown in Figure 3.20. The lower loss is highly desirable for applications in this dissertation.

Figure 3.19: Material properties extracted from s-parameter simulation of a $5mm$ unit cell with various shunt inductance values.

Figure 3.20: Loss extracted from s-parameter simulation of a $5mm$ unit cell with various shunt inductance values.

Another analysis is done for varying the size of the unit cell while maintaining a shunt inductance of $1nH$. This results in a more negative dielectric as shown in Figure 3.21. The larger negative dielectric constant is beneficial to confine EM field. In addition, a larger negative value means the fields will not penetrate as far into the unit cell and thus results in lower losses. The loss and magnetic constant remained unchanged with different unit cell sizes.

Figure 3.21: Effect of unit cell size on dielectric constant. The shunt inductor has a value of $1nH$.

Additional simulations are performed for a frequency of $1GHz$ and other parameters are varied. The next simulation performed is for a shunt inductance ranging

from $0.01 - 10\text{ nH}$. In Figure 3.22a it can be seen that there is an optimal inductance value to achieve a large negative dielectric value. However, the largest inductance value retains a very negative dielectric value and is useful for confining EM energy. The magnetic constant can be seen to vary insignificantly with increase in inductance as seen in Figure 3.22b. This is because the unit cell is small and the frequency is low enough that the shunt inductance has little affect on the magnetic constant. If the frequency or cell size are larger the shunt inductance would affect the magnetic constant as seen in Figure 3.19b. As shown previously, the loss decreases as the inductance increases and can be seen in Figure 3.23.

Figure 3.22: Material properties extracted from s-parameter simulation of 1 mm unit cell with value of shunt inductor swept.

Figure 3.23: Loss tangents extracted from s-parameter simulation of unit cells with value of shunt inductor swept.

The next analysis is done for a $1mm$ unit cell with a shunt inductance of $1nH$, resistor in the shunt configuration, and the series inductance value is swept. It can be seen in Figure 3.24a that increasing series inductance results in a lower negative dielectric value, while increasing the magnetic constant as shown in Figure 3.24b. Figure 3.25 demonstrates that increasing series inductance results in lower losses.

Figure 3.24: Material properties extracted from s-parameter simulation of $1mm$ unit cells with $1nH$ shunt inductance and sweeping series inductance.

Figure 3.25: Loss tangents extracted from s-parameter simulation of $1mm$ unit cells with $1nH$ shunt inductance and sweeping series inductance.

The last simulation is performed for a $1mm$ unit cell with a $1nH$ shunt inductor, the resistor in the shunt configuration, and the plate separation is swept. The results in Figure 3.26a show that a larger plate separation results in a more negative dielectric. The results in Figure 3.26b shows that the magnetic constant and dielectric constant are linked. For this sweep if there was no link the magnetic constant would have been unity. The increase in plate separation results in a decrease in loss as seen in Figure 3.27.

Figure 3.26: Material properties extracted from s-parameter simulation of $1mm$ unit cell with $1nH$ shunt inductance and sweeping value of place separation.

Figure 3.27: Loss tangents extracted from s-parameter simulation of $1mm$ unit cell with $1nH$ shunt inductance and sweeping value of place separation.

The series and shunt components affect both the permeability and permittivity, while the series component affects both permittivity and permeability the most. The method used previously to derive equations for the effective dielectric constant and dielectric loss tangent assume that the shunt component only affects permittivity and the series component only affects permeability. Since this is not the case, deriving the equations this way does not yield as accurate results, however the results are still useful. The shunt inductance (L_1) with a resistor (R_1) in series was transformed into

a negative capacitance in parallel with a resistor. The dielectric constant is given by Equation 3.16 and the dielectric loss tangent is given by Equation 3.17. As stated previously, the negative value for the dielectric loss tangent does correspond to loss and not gain. The equation for dielectric loss tangent shows the behavior observed in previous analyses. The main way to decrease loss is to decrease the resistance of the switch or increase the shunt inductance.

$$\epsilon' = -\frac{L_1 d}{(L_1^2 \omega^2 + R_1^2) W^2} \quad (3.16)$$

$$\tan \delta = -\frac{R_1}{\omega L_1} \quad (3.17)$$

where:

- L_1 is the shunt inductor used in the transmission line equivalent circuit to realize a negative dielectric.
- R_1 is the resistor in series with the shunt Inductor.
- d is the parallel plate separation.
- W is the width of the parallel plate unit cell.

The negative permittivity unit cell shows significantly higher losses than the double positive unit cell. This is an issue that appears difficult to remedy. The FPMSs in this work makes use of the negative permittivity material to realize programmable waveguides with a positive dielectric core sandwiched between negative dielectric material. This type of guide does not result in the EM energy interacting significantly with the negative dielectric in most cases. If the losses prove too large in some applications it may be advantageous to implement waveguides using the more traditional slab waveguides that consist of a positive dielectric core sandwiched between

two lower positive dielectric constant materials.

3.3 Implementation of the waveguide

The loss from the negative permittivity material seems prohibitively high, however this is not the case. A waveguide composed of a positive permittivity core sandwiched between a negative permittivity material is the same form as the slab waveguide. In this configuration the EM energy does not interact strongly with the negative dielectric. The slab waveguide equations do not make any assumptions on the sign of material properties and are valid to use for analyzing this type of waveguide. Equations 3.18-3.20 are the slab waveguide equations required to calculate the propagation constant for the first mode of the guide [140]. Equation 3.18 must be solved numerically for k_x . Given loss in the material, solving for k_x is more involved because one must solve for the real and imaginary parts of k_x simultaneously.

$$\tan k_x = \sqrt{\frac{\omega^2(\mu_{core}\epsilon_{core} - \mu_{side}\epsilon_{side})}{(k_x W)^2} - 1} \quad (3.18)$$

$$k = \omega \sqrt{\mu_{core}\epsilon_{core}} \quad (3.19)$$

$$k_z = \sqrt{k^2 - k_x^2} \quad (3.20)$$

where:

- k_z is the complex propagation constant of the waveguide.
- ω is the frequency.
- μ_{core} and ϵ_{core} is the permeability and permittivity core waveguide material.
- μ_{side} and ϵ_{side} is the permeability and permittivity of the material on sides of the

guide.

- $2W$ is the width of the waveguide core.

To demonstrate that reasonable losses are obtainable, unit cell properties analyzed previously are used in the calculations of slab waveguides. The first guide analyzed uses a positive permittivity unit cell with a width of $1mm$ and is filled with a substrate that has a dielectric constant of 2. The properties of this material were given previously in Figures 3.5 and 3.6. The side of the waveguide consists of a negative permittivity $1mm$ wide unit cell that uses a shunt inductance of $1nH$ in series with a 5Ω resistor and the material properties are given in Figures 3.28 and 3.29. A waveguide implemented with these materials yields the propagation characteristics shown in Figure 3.30. The core of the waveguide is $60mm$ wide and the sides are assumed to extend to infinity. The propagation constant is similar to that of SIW and rectangular waveguides. The attenuation constant shows a relatively low loss of less than $0.2dB/\lambda_g$ and the behavior close to cutoff is the same as that observed with SIW and rectangular waveguides. This demonstrates that given an extremely lossy sidewall, it is possible for only a modest amount of loss will be incurred.

Figure 3.28: Material properties extracted from s-parameter simulation of $1mm$ wide and $1mm$ thick unit cell with a $1nH$ shunt inductance in series with a 5Ω resistor.

Figure 3.29: Loss tangents extracted from s-parameter simulation of $1mm$ wide and $1mm$ thick unit cell with a $1nH$ shunt inductance in series with a 5Ω resistor.

Figure 3.30: Propagation characteristics of a 60mm wide slab waveguide. (a) is the propagation constant and (b) is the attenuation.

Further analysis will be done using lossless and dispersion-less material for the core of the waveguide and lossy material for the sidewalls. The purpose of this analysis is to determine how to further minimize the losses from the sidewalls. The next analysis is the same as the previous one, except the core material is replaced with a dispersion-less material with a dielectric constant of 2, 4.9, or 5. The propagation characteristics of these waveguides are given in Figure 3.31. These results indicate that a larger dielectric material for the core will help minimize the loss contribution from the sidewalls.

Figure 3.31: Propagation characteristics of a 60mm wide slab waveguide with various core dielectric materials. (a) is the propagation constant and (b) is the attenuation.

The next configuration analyzed is similar to the previous one. This one uses a core with a dielectric constant of 2 and the properties of the sidewall are varied. For the sidewall, a unit cell consisting of a shunt inductance of $1nH$ in series with a 5Ω resistor is used. Sidewall unit cell sizes of 1, 2, or 5mm are used and the negative dielectric constants for the cells were shown previously in Figure 3.21. The smaller unit cell yields a more negative dielectric constant, while the magnetic constant and loss of this material are the same for all three unit cell sizes. The calculated properties of the waveguide are shown in Figure 3.32. From the results it can be seen that a less negative dielectric constant for the sidewalls lowers the cutoff frequency. This is because the EM energy penetrates further into the sidewall, thus effectively making the waveguide wider. The EM energy penetrating further into the sidewall also results in increased loss.

Figure 3.32: Propagation characteristics of a 60mm wide slab waveguide with various negative dielectric constants for the sidewall. The plot labeled 1mm is the most negative permittivity and the plot labeled 5mm is the least negative permittivity. (a) is the propagation constant and (b) is the attenuation.

It was shown that a higher dielectric constant for the core and a more negative dielectric constant for the sidewalls of the waveguide leads to lower loss. The larger dielectric constant for the core material and the more negative dielectric constant for the sidewall can be achieved with a smaller unit cell size.

3.4 Conclusions

This chapter uses a basic transmission line unit cell equivalent circuit to analyze placement of a lossy component and determine methods to mitigate the effects of the lossy component. It is interesting to note that many of the characteristics that yield a well performing positive dielectric also yield a well performing negative dielectric. Engineering a positive dielectric material will yield the best performance by adhering to the following rules:

- Unit cell must be less than $\lambda/8$ in electrical size.

- Lossy component should be in the shunt configuration.
- Shunt capacitance should be as small as possible to minimize losses from shunt lossy component.
- Series inductance should be as large as possible to minimize losses from any series resistance.
- Increased plate separation lowers losses by decreasing capacitance and increasing inductance.

Design of a negative dielectric material will yield best results by following these rules:

- Unit cell must be less than $\lambda/8$ in electrical size.
- Minimize cell size to maximize negative dielectric value.
- Maximize shunt inductance value to minimize loss from shunt lossy component. There is a trade off with negative dielectric value.
- Maximize series inductance value to minimize loss. There is a trade off with maximizing negative dielectric value.
- Maximize plate separation to minimize loss and increase negative dielectric value.

It was shown that relatively low loss slab waveguides can be implemented with these materials by using a large dielectric constant core and a very negative dielectric constant sidewall. The next chapters make use of this information to implement a field programmable microwave substrate (FPMS).

Chapter 4

Field Programmable Microwave Substrate

4.1 Introduction

Realizing a programmable microwave circuit that is as versatile and powerful as an FPGA requires the ability to implement programmable waveguides. A method of implementing low loss programmable waveguides is shown in Chapter 3. This chapter makes use of the material presented in Chapter 3 to realize prototype microwave circuits with unprecedented programmability.

This chapter discusses two different implementations of the FPMS. The first device is implemented using an FR4 substrate and off-the-shelf components. The section on this implementation presents the unit cell used and the effective material properties of the unit cell. The FR4 implementation is used to demonstrate programmable waveguides, amplifiers, and oscillators. Section 4.3 presents an FPMS implemented using LTCC and custom designed CMOS chips. This implementation is significantly smaller than the FR4 implementation and functions over a similar frequency range. Various waveguides are implemented using this FPMS.

4.2 FPMS Implemented on FR4 Substrate

The first FPMS presented is one that is fabricated using industry standard FR4 substrate, off-the-shelf components, and a pick-and-place service. This section describes the unit cell used to realize the FPMS and demonstrate the capabilities of an FPMS implemented using this unit cell. The capabilities include waveguides, amplifiers, and oscillators functioning over various frequency bands.

4.2.1 Unit Cell

The unit cell implemented on an FR4 substrate is shown in Figure 4.1. This unit cell is designed to switch the effective dielectric constant between positive and negative values. This is achieved using a shunt varactor with a large tuning range and large inductive package parasitics ($1.8nH$ for the varactor chosen). This allows for the varactor to be switched between behaving as a capacitor or an inductor over a large range. For the varactor chosen, no bias yields a $12pF$ capacitance and is series resonant at about $1GHz$. Biased at $25V$ the varactor has a capacitance of $0.75pF$ and is series resonant at $4.5GHz$. It is ideal to use a low loss switch rather than a varactor because a much larger range of frequencies can be covered with a single circuit. However, it is difficult to find an off-the-shelf switch with sufficient performance.

Figure 4.1: Unit cell used to construct FPMS. The components used are: V1 is a Infineon varactor (BB 833 E6327); R1 is a $10k\Omega$ resistor (ERJ-2GEJ103X); C1 is a $2200pF$ decoupling capacitor (LLL185R71H222MA01L); and the metallization in brown. The metal cross seen has $20mil$ wide lines (W) and is $200 \times 200mil$ ($L = 100mil$). Spacing from top metal to ground is $46.1mil$ and the dielectric is FR4 with a dielectric constant of 4.9 and dielectric loss tangent of 0.02. The via for biasing the varactor goes through the ground plane with a $10mil$ clearance. The terminations shown have an impedance of 88.84Ω .

ADS Momentum is used to simulate the metal structuring for the unit cell and a co-simulation (EM and circuit simulation combined) is performed with the unit cell shown in Figure 4.1. This is done for various biases on the varactor and the effective material properties are shown in Figures 4.2-4.6. No measurements are made on fabricated unit cells to extract the properties. The properties shown in Figure 4.2 demonstrates that the effective dielectric constant of the unit cell can be changed with the bias on the varactor. As expected the self-resonance frequency of the unit cell is affected by changing the capacitance of the varactor. The dielectric loss tangent

of the unit cell is shown in Figure 4.3 and shows a rapidly increasing loss close to resonance. This demonstrates it is best to bias the varactor so the resonance point is sufficiently far from any frequency of interest. The magnetic constant shown in Figure 4.4 is larger than unity due to an increased series inductance from the metal cross structure used. The decrease in magnetic constant with frequency is similar to that shown in Chapter 3. The magnetic loss tangent shown in Figure 4.4 shows a non-zero value due to metal losses in the metal cross. The value decreases below zero as frequency increases and is reported in the literature as mentioned previously. Adding magnetic and dielectric loss tangent results in an overall loss in power.

Figure 4.2: Effective dielectric constant of unit cell.

Figure 4.3: Effective dielectric loss tangent of unit cell.

Figure 4.4: Effective magnetic constant of unit cell.

Figure 4.5: Effective magnetic loss tangent of unit cell.

The properties of the negative dielectric unit cell are shown in Figure 4.6. This is achieved with no bias on the varactor. The losses are relatively high, but in the slab waveguide most of the EM energy is confined in the positive dielectric material so this is less of a problem. The positive dielectric constant at frequencies lower than about 1GHz is because the varactor acts as a capacitor at those frequencies and above 1GHz it behaves as an inductor.

Figure 4.6: Effective properties of unit cell when biased at 0V.

4.2.2 FPMS

A 16×16 element FPMS is implemented using the previously described unit cell and is shown in Figure 4.7 and Figure 4.8. The backside is the routing for biasing the varactors and the signals for the serial to parallel converters with voltage buffered outputs (HV57908PG-G). To connect to the FPMS a microstrip to SIW transition is used. The transition chosen is much larger than required, but it is reliable, broadband, and size is of no concern. A close up of the 256 unit cells is shown in Figure 4.9. On the top and bottom right hand side inductors can be seen (circled in red) that ground the anode of the varactors. The FPMS implemented on FR4 consists of three microwave ports, a transistor in the common emitter configuration, and 256 programmable unit cells. The components circled in blue on the top of Figure 4.9 are two 3pF capacitors that are used as DC blocks and to filter out low frequency signals. Without the capacitors the circuit oscillates in the hundreds of MHz range. The layout for the transistor connected to the FPMS is shown in Figure 4.10. The bill of material for the FPMS is given in Table 4.1.

Figure 4.7: Picture of front of board used to demonstrate an implementation of the FPMS on FR4. The board measures 12.38x11.56 inches.

Figure 4.8: Picture of back of board used to demonstrate an implementation of the FPMS on FR4. The board measures 12.38x11.56 inches. The traces seen are mainly for biasing the varactors.

Figure 4.9: Close up picture of FPMS which is 3.2x3.2 inches and consists of 256 unit cells.

Figure 4.10: Layout of transistor and circuitry for connecting to the FPMS. The width of the lines are 80mil. The base of the transistor is connected on the left and is biased through the inductors (L1). The transistor (T1) is biased with a collector current of 70mA and collector-emitter voltage of 2V. The length of the line on the left is 250mil and the length of the line on the right is 175mil. The capacitors labeled C2 are a 3pF capacitor (Murata GJM1555C1H3R0CB01D).

Item #	Ref Des	Description	Qty	Manufacturer	Mfg Part #	Package	Type
1	T1	NPN bipolar RF transistor	1	Infineon	BFP 450 H6327	SOT-343	SMD
2	C1	2200pF Decoupling Capacitor	258	Murata	LLL185R71H222MA01L	0306	SMD
3	R1	Thick film 10K resistor	256	Panasonic	ERJ-2GEJ103X	0402	SMD
4	L1	Wirewound RF choke inductor	5	Murata	LQW15AN36NJ00D	0402	SMD
5	QPF1	High Voltage shift register	4	Microchip technology	HV57908PG-G	QFP80 (14x20)	SMD
6	V1	Varactor	256	Infineon	BB 833 E6327	SOD-323	SMD

Table 4.1: Bill of materials.

The FPMS is programmed with a bit stream sent to the HV57908PG-G chips from an FPGA. Matlab is used to communicate with the FPGA and the power supply. The power supply controls the bias on all varactors. Measurements of the FPMS are done with an HP 8595A spectrum analyzer and an Agilent 8722ES network analyzer. The network analyzer is calibrated using an HP 85052D SOLT calibration kit with the network analyzer using a sweep power of -10dBm. There is no structure fabricated to measure the back to back characteristics of the transition and for this reason the simulated results shown in Figure 4.11 are used to subtract out the losses from the transition. The next part of this section demonstrates various circuits implemented using the FPMS. Due to the nature of the FPMS an exhaustive demonstration of all possible circuits is simply not practical.

Figure 4.11: Insertion loss of back-to-back transition simulated using ADS Momentum.

Three different waveguides are shown in Figure 4.12. The blue squares represent a positive dielectric while the white squares represent a negative dielectric constant. The three external ports are the grey boxes on the left, right, and bottom. The connections for the transistor are on the top, the left side is the collector connection, and the right side is the base connection.

Simulations of the FPMS fabricated on FR4 are performed in ADS. All the metalization is simulated in ADS Momentum. This is done in three separate simulations: one simulation is done for the transition; one for the unit cell; and one for the metal required for connecting the transistor to the FPMS. A simulation is then ran using the EM simulation data for the metal and models provided from the component manufacturers to simulate the full FPMS.

The waveguide shown in Figure 4.12a is simulated in ADS and the simulated and measured results are shown in Figure 4.13. It can be seen that the simulated and measured results show a decent match. The differences between the results could be due to multiple reasons such as: component tolerances; fabrication tolerances; imprecise material values; and accuracy of EM simulations. The measured results show the return loss is not symmetrical. There are two possible reasons, the first

could be due to damage on port 1 of the board. The board was accidentally bent during initial testing which caused visible damage on the top of the board. The visible damage has negligible effect on the circuit, however the ground plane which is not visible may be damaged and could affect the circuit. Alternatively, the lack of symmetry could be due to the non-symmetrical component filling the transitions as seen in Figure 4.9. The waveguide exhibits a bandpass characteristic. The low frequency cutoff is expected since the guide functions similarly to the SIW. The high frequency cutoff is due to the series resonance of the unit cell. The insertion loss of the waveguide at $2GHz$ is $2.3dB$. For comparison, a 50Ω microstrip line on the same substrate would have an insertion loss of $0.55dB$. The extra loss is partially attributed to the lossy negative dielectric that make up the side walls of the guide. In addition, almost all the EM fields are contained within the substrate, which is not the case with a microstrip line. It should be noted that the programmable waveguide is electrically longer than the microstrip line it is being compared to.

The measured results shown in Figure 4.14 are for the waveguides shown in Figure 4.12b and 4.12c. It can be seen that the wider waveguide has a lower cutoff frequency. The results shown in Figure 4.14b shows a lower high frequency cutoff because the positive part of the guide is biased at $10.72V$ rather than $25V$. This results in the varactor having a larger capacitance and the unit cell having a lower frequency resonance. All the waveguide measurements are made with the transistor biased, and it was seen that the transistor did not affect the waveguides.

Figure 4.12: Waveguides implemented using the FPMS for various frequency ranges.

Figure 4.13: Comparison of simulated and measured waveguide shown in Figure 4.12a

Figure 4.14: Measured waveguides of different widths.

The simulated propagation characteristics for the waveguides are given in Figure 4.15.

Figure 4.15: Waveguide propagation characteristics calculated using slab waveguide equations and effective materials properties.

The propagation constants shown in Figure 4.15a demonstrate a distinctive cutoff frequency for the waveguides that are 4 and 6 unit cells wide. However, it can be seen that for frequencies below the cutoff frequency the propagation constant increases. This is because the dielectric constant of the sidewall changes from negative

to positive at lower frequencies as seen in Figure 4.6. The 12 unit cell wide waveguide does not show a distinctive cutoff frequency for the same reason. All three waveguides show a rapidly increasing propagation constant at higher frequencies. This is because the effective material properties for the waveguide core increase rapidly at these frequencies as seen in Figure 4.2. The attenuation for the waveguides is shown in Figure 4.15b. At low frequencies the loss increases rapidly as normally seen with SIW and rectangular waveguides. At higher frequencies the loss increases rapidly due to the unit cell being self-resonant and this matches the loss tangents for the waveguide core material shown in Figures 4.3 and 4.5.

Three different amplifier implementations are shown in Figure 4.16. The amplifiers are designed by utilizing the programmability of the FPMS. First the signals are routed from the ports to the transistor using waveguides. The width of these waveguides are set depending on the desired center frequency of the amplifier. The s-parameters are displayed on a VNA to determine if the waveguides are sized correctly. Next, stubs are added and adjusted. Finally, the bias on the varactors are adjusted to fine tune the frequency response of the amplifier. The measured and simulated results for the configuration shown in Figure 4.16a are given in Figure 4.17. The results show a good match between simulation and measurement. The peak around 2 GHz that is larger in measurement than simulation is likely due to a non-linear feedback effect that is not accounted for in the simulation. The amplifier utilizes a stub match that also facilitates feed back to reduce S_{12} . The single row of unit cells between the stubs allows for EM energy to be coupled between the stubs, this is because the effective dielectric constant of the material at 1.5GHz is -28 . Measured results for amplifiers centered at 1.05GHz and 2GHz are shown in Figure 4.17 and correspond to the configurations shown in Figure 4.16b and 4.16c, respectively.

(a) $f_c = 1.5GHz$, Bias=25V (b) $f_c = 1.05GHz$, Bias=8V. (c) $f_c = 2GHz$, Bias=25V

Figure 4.16: Amplifiers implemented with the FPMS.

Figure 4.17: Comparison of measured and simulated s-parameters for 1.5GHz amplifier shown in Figure 4.16a. Gain and 3dB bandwidth are 10.7dB and 140MHz, respectively.

Figure 4.18: S-parameters for measured amplifiers. The amplifier in (a) is centered at 1.05GHz with a gain of 9.2dB and 3dB bandwidth of 110MHz. The amplifier in (b) is centered at 2GHz with a gain of 5dB and a 3dB bandwidth of 230MHz.

The last circuits to be shown are oscillators. The oscillators are designed in a similar manner as the amplifiers. Waveguides are used to give positive feedback between the base and collector of the transistor. The oscillation of the circuit is analyzed by connecting one port to a spectrum analyzer and a waveguide is programmed to connect close to the feedback waveguide without making contact. Next, the length and shape of the feedback waveguide are modified to achieve oscillation close to the desired frequency. The waveguide that connects to the port is then modified to optimize power coupled out. This may require further modification of the feedback waveguide. Precise oscillation frequency is achieved by tuning the dielectric constant of the waveguides. The measurements for the oscillators are not made in a Faraday cage. Spectrum analyzer measurements with no bias on the transistor shows signals being received by the FPMS with power levels around -70dBm at 0.875GHz and 3.5GHz. It is quite possible that other signals below the noise floor of the spectrum analyzer (-80dBm) are also being received by the FPMS. Three different configurations for oscillators are shown in Figure 4.19. Each of the oscillators achieve oscillation via

positive feedback. This can result in very large gain for signals that are below the noise floor for the initial measurements without any power. The measured results shown in Figure 4.20 are for the oscillator shown in Figure 4.19a. Harmonic balance simulations are performed in ADS for the same configuration yielding an oscillation of $-1.92dBm$ at $2.584GHz$. The measured results show an oscillation of $3.61dBm$ at $2.5GHz$. This is a 3.4% difference in oscillation frequency and a $5.53dB$ difference in power. The difference in oscillation frequency is not surprising since resonances are very sensitive to small changes. The larger difference in power is attributed to the higher gain of the transistor and lower losses in the circuit at the lower frequency of oscillation. The larger spurs seen in Figure 4.20 are attributed to signals present in the measurement environment. The measurement results for a 1.2 GHz and a 1.5GHz oscillator are shown in Figure 4.21. Harmonics can be seen for the 1.2GHz and 1.5GHz oscillators.

The configuration shown in Figure 4.19c demonstrates that a programmable microwave circuit can realize conventional circuit performance with non-conventional configurations, such as a smiley face.

Figure 4.19: Oscillators implemented with the FPMS.

Figure 4.20: Oscillators implemented with the FPMS. 3.61dBm at 2.5GHz

Figure 4.21: Oscillators implemented with the FPMS. (a) is oscillator configuration shown in Figure 4.19b and (b) is the configuration shown in Figure 4.19c.

This section presents an FPMS implemented on FR4 that is used to realize a variety of microwave circuits such as waveguides, amplifiers, and oscillators. The next section presents a more miniaturized version of the FPMS.

4.3 FPMS Implemented on LTCC

FR4 does not make the best substrate for microwave applications and it would be advantageous to use a material with superior characteristics such as LTCC. The LTCC

material used is DuPont 9K7 which has a dielectric constant of 7.1 and a dielectric loss tangent of 0.001. The metal is $10\mu m$ thick gold and the other important design characteristics are given in Table 4.2.

	Dimension (μm)
Minimum Line Width	125
Minimum Line Spacing	150
Minimum Line to Via Spacing	200
Minimum Via Diameter	150
Minimum Via Pitch	$2.5 \times$ Via diameter

Table 4.2: Key LTCC process design limitations

This section presents an FPMS implemented on LTCC. The FPMS measures $3 \times 2 cm$ and functions from $2.5 - 3.5 GHz$.

4.3.1 Unit Cell

The unit cell presented in the last section is mainly designed using an equivalent circuit. The design of the unit cell in this section is performed mainly based on electromagnetic simulation. The basic unit cell is shown in Figure 4.22 and looks very similar to the loaded posts used to realize tunable SIW circuits [132–136]. The effective material properties of the unit cell are modified by attaching components between the two posts. The simplest cases are: open circuited for a positive dielectric constant; and short circuited for a negative dielectric constant. The electric field patterns of a open-circuited unit cell are shown in Figure 4.24. Some of the strongest

fields are between the post and the top plate, and between the two posts. Simulations of the unit cell are performed in HFSS by exciting the cell with a plane wave. This is done by extending an air box at least $\lambda/4$ from the structure and applying a perfect electric conductor (PEC) to the top and bottom and a perfect magnetic conductor (PMC) to the sides as shown in Figure 4.23. The air box in the center above the unit cell has an absorbing boundary condition applied to the top and PEC applied to the sides not marked PMC. This structure is excited with the electric field of the wave normal to the PEC. The effect of the air boxes needed to realize a plane wave are de-embedded from the s-parameters using HFSS. The s-parameters from these simulations are used in the NRW algorithm to calculate the effective material properties. The accuracy of the extracted materials can be increased by accounting for boundary effects between the air box and the artificial material [141]. Accounting for boundary effects when determining material properties is not done for this work. One reason is because material properties extracted using more than one unit cell show little variation from properties extracted using one cell. In addition, the unit cell is used in a programmable structure and minor variation from the expected properties are easily mitigated.

Figure 4.22: Post unit cell

Figure 4.23: Simulation setup for LTCC unit cell.

Figure 4.24: The electric fields in the open-circuited post unit cell excited by a plane wave.

The unit cell shown previously is a more idealized structure, and the unit cell implemented in LTCC is the cell shown in Figure 4.25a. The top and bottom plate are

interior to the LTCC tile and are meshed to prevent delamination. The unit cell is $1.5 \times 1.5 \text{ mm}$ and the distance between the meshed plates is 0.448 mm . The post extends up 0.224 mm above the top metal meshed plate. Twelve of the unit cells are used to realize a master unit cell as shown in Figure 4.25b. The center post in the LTCC master cell is RF ground. There are DC power and ground lines, and two signal lines to communicate with the CMOS chip. The master unit cell has a CMOS chip flip-chip bonded to it. The master unit cell is rectangular rather than square to allow for a single CMOS chip to control twelve unit cells rather than nine.

Figure 4.25: (a) is the cell that makes up the FPMS in LTCC, and (b) is a master unit cell consisting of twelve unit cells.

The CMOS circuitry that connects to each unit cell of the master cell is shown in Figure 4.26. This is the circuit used for simulations, however the CMOS chip has a slightly different layout. On the CMOS chip each switch circuit that connects to the LTCC posts has an input pad, but every switch circuit shares the same output pad. In the switch circuit, double diodes and 10 pF capacitors can be seen which are used to meet the requirements for electrostatic discharge (ESD) protection. The double diodes add significant parasitic capacitance to ground. It is preferable to remove both the double diodes and series capacitors from the design, but this increases the

risk of non-functional circuits. The switch is an NFET with a finger width of $60\mu m$ and length of $120nm$. The varactor is a differential NMOS varactor with a tuning range from $58 - 152fF$. NMOS varactors are lossier than the hyperabrupt varactor available in the process but the NMOS one is chosen because the voltages were compatible with the digital logic used. Control of the circuit shown in Figure 4.26 is achieved with a digital circuit that can control each of the twelve unit cells individually. The capacitor parallel to the varactor was seen to improve loss of the unit cell at the cost of lowering tuning range of the varactor. The digital logic is implemented using ARM standard cells for the process.

The simulated input resistance and capacitance for the switch is shown in Figure 4.27. The capacitance when the switch is off can be changed from 0.83-0.86pF using the varactor. This limited ability to tune the input capacitance of the circuit is not used and the performance of the circuit would have been improved if it was omitted. The only measurements performed on the CMOS chip are to confirm that it functions and the high frequency characteristics are not measured. Simulation data is relied upon for two reasons. First, research groups in the Department of Electronics at Carleton University make extensive use of the CMOS process with reliable results that matched well with simulations. Second, measuring the characteristics can not be performed directly on the CMOS chip and would require the chips to be flip-chip bonded to an additional substrate for characterization. This additional effort is deemed unimportant for demonstration of a functional prototype.

Figure 4.26: Circuit connected to each unit cell implemented on the CMOS chip.

Figure 4.27: Input impedance of CMOS switch circuit with the switch in the on- and off-states. These results are for an equivalent circuit with a resistor in series with a capacitor.

The master unit cell is simulated in HFSS using the same method as described

earlier for a single unit cell. Lumped ports are used where the CMOS chip is connected. The s-parameters for the master unit cell are used in ADS in conjunction with the s-parameters for the switch circuit in either the on-state or off-state. The s-parameters for the entire circuit (LTCC master unit cell and CMOS switch circuit) are used in the NRW algorithm to calculate the effective material properties. The effective material properties for the switch in the off-state are shown in Figures 4.28 and 4.29. The master unit cell is characterized with the unit cell in two orientations, this gives material lengths of 6 and 4.5mm. The NRW algorithm produces less accurate results when the electrical length of the material being characterized is close to 180° . This is the case for characterizing the long side (6mm) of the master unit cell at about 3.3GHz and for the short side (4.5mm) at about 4GHz. Extracting the effective properties of the master unit cell in both orientations (i.e. long and short) helps mitigate this issue by using the properties of the other orientation when close to an electrical length of 180° . The material properties for both orientations shows the longer orientation has a slightly lower resonant frequency. This is likely because the metal lines connecting the LTCC posts to the CMOS chip do not have rotational symmetry.

The dielectric and magnetic constant shown in Figure 4.28 demonstrate that the unit cell is self-resonant at about 4.5GHz. The metal lines used to connect the CMOS chip to the LTCC tile result in a significant lowering of resonant frequency of the unit cell. Removing these lines and having the switch closer to the post would be highly beneficial. The loss tangent shown in Figure 4.29 is 0.028 at 1GHz. For comparison the effective dielectric and magnetic loss tangent of the unloaded (i.e. nothing connected) master unit cell are shown in Figure 4.30. The magnetic loss tangent of the unloaded cell is about 0.02, which shows that a significant amount of the loss is from the metal that makes up the parallel plates. The primary method that can be used to

reduce this loss in LTCC is to increase the spacing between the parallel plates. The LTCC design is constrained by a predefined stack-up that prevents this from being implemented. The dielectric loss contribution can be lowered by decreasing the input capacitance of the CMOS switch circuit. The double diodes used for ESD protection contribute significantly to this input capacitance and removal of these is beneficial.

Figure 4.28: Effective material properties of master unit cell with switch off. The plot labeled 6mm are the material properties for the long side and the plot labeled 4.5mm are for the material properties for the short side.

Figure 4.29: Effective material properties of master unit cell with switch off. The plot labeled 6mm are the material properties for the long side and the plot labeled 4.5mm are for the material properties for the short side.

Figure 4.30: Loss of unloaded master unit cell

The simulated effective properties of the master unit cell with the CMOS switch in the on-state are given in Figures 4.31-4.32. The dielectric constant is positive at low frequency and becomes negative around 2GHz . This is because of the 10pF capacitors used on the input and output of the CMOS switch circuit for ESD reasons. The switch in the on-state has a much larger resistance than state-of-the-art switches and this resistance value could easily be reduced. The dielectric constant is not as negative as desired and this should be made more negative to decrease losses.

Figure 4.31: Effective material properties of master unit cell with switch off. The plot labeled 6mm are the material properties for the long side and the plot labeled 4mm are for the material properties for the short side.

Figure 4.32: Effective material properties of master unit cell with switch on. The plot labeled 6mm are the material properties for the long side and the plot labeled 4mm are for the material properties for the short side.

4.3.2 FPMS

A picture of the FPMS implemented using LTCC is shown in Figure 4.33 and the metal structuring used in HFSS simulations is shown in Figure 4.34. The top 2/3 of the tile is the FPMS and the bottom 1/3 is a similar circuit, but can be manually programmed by soldering on surface mount components; this is a backup circuit and is not used. A representation of the FPMS is shown in Figure 4.35. Each square represents one programmable unit cell and the positioning of the CPW-SIW transitions are shown at each port. The transition is positioned between the first and second unit cells. This type of transition is commonly used in SIW circuits [129]. A close up of the transition from simulations in HFSS is shown in Figure 4.36. The transition should be positioned so the posts normally used to realize the unit cells will load the CPW center conductor. Unfortunately, a mistake made during layout results in the transition not positioned optimally and thus results in a poor return loss. The edge of the FPMS is surrounded with vias to help minimize losses by not requiring programmable negative dielectric for some of the sidewalls.

Figure 4.33: Picture of the LTCC tile with CMOS chips flip chip bonded to it. The FPMS is on the top part of the tile and the circuit is 3cm x 2cm. The non-functional CMOS chips are marked with a red X.

Figure 4.34: LTCC tile metal structure used for simulations in HFSS. View is shown from the other side as shown in Figure 4.33.

Figure 4.35: LTCC tile layout with ports and transition placement shown.

Figure 4.36: 90° bend CPW-SIW Transition.

The LTCC tile is connectorized using Murata SWG connectors and measured using an Agilent 8722ES network analyzer. The CMOS chips are programmed using Matlab to send commands to an FPGA that generates the appropriate bit stream to program the CMOS chips. All unit cells are programmed for a positive dielectric constant and each unit cell is tested by switching it to a negative dielectric constant and measuring the s-parameters for ports 1 and 2 connected. Equation 4.1 is used to generate Figure 4.37. The number 250 is chosen so the equation yields numbers from 0 – 1, which is used to show in grey scale how much each unit cell affects the circuit.

$$\sum_{freq=1GHz}^{5GHz} (|S_{11}a - S_{11}b| + |S_{22}a - S_{22}b| + |S_{21}a - S_{21}b|)/250 \quad (4.1)$$

where,

- $S_{11}a$, $S_{22}a$, and $S_{21}a$ are s-parameter measurements with all the unit cells programmed with a positive dielectric constant.
- $S_{11}b$, $S_{22}b$, and $S_{21}b$ are s-parameter measurements with a single unit cell under test programmed with a negative dielectric constant.

It is determined that three of the CMOS chips do not respond to programming and only one of the three powers on. The CMOS chips that power on have an initial state with the switches all in the on-state. All three of the chips are connected by port 3 and the 36 unit cells in black on the top are non-functional. If all the unit cells function the same, it is expected that the center of the waveguide (4.37) is close to white and the sides are darker. However, it can be seen that some cells do not respond to programming or have much less affect on the s-parameters than they should. This is either due to non-functional chips or poor flip-chip connections.

Figure 4.37: Measurement of effect each cell has on s-parameter measurements. White represents a unit cell, when switched negative, that has the largest effect on the s-parameters of a thru waveguide, while black represents a unit cell that has a negligible effect on the s-parameters.

Waveguides between ports 1 and 2 are measured and are shown in Figures 4.38-4.40. The width of the waveguide is varied by progressively turning the top most row of unit cells from positive to negative. The bottom four rows are always positive. The results demonstrate the ability to change the lower cutoff frequency by changing the width of the waveguide. The upper cutoff frequency is due to the series resonance of the unit cell. As the width of the waveguide is reduced the loss of the waveguide increases because the electromagnetic energy interacts more strongly with the lossy sidewalls. Figure 4.40b shows the results for all unit cells switched to a negative dielectric constant and demonstrates decent isolation.

Figure 4.38: S-parameters of Waveguides implemented on LTCC based FPMS.

Figure 4.39: S-parameters of waveguides implemented on LTCC based FPMS.

Figure 4.40: S-parameters of Waveguides implemented on LTCC based FPMS.

Figure 4.41: Comparison of insertion loss for LTCC FPMS thru waveguides.

A comparison of a simulated waveguide with a width of six unit cells and a measured waveguide is shown in Figure 4.42. The insertion loss and shape of the measured and simulated guides are similar. The higher cutoff frequency of the simulated waveguide is due to a few factors. The parasitics from the flip-chip connections are

not included in the simulation. The main parasitic from these add inductance that lowers the upper frequency cutoff of the waveguide. The modeling of the CMOS circuitry is extracted from simulation of the one switch circuit rather than extracting the entire CMOS chip. Accuracy of the simulation can be increased by accounting for all these effects, but the results are sufficiently accurate and increasing the accuracy does not benefit the discussion.

Figure 4.42: Comparison of measured and simulated waveguide that is six unit cells wide.

4.4 Conclusion

This chapter presents two implementations of the FPMS. One implementation uses FR4 and off-the-shelf components. This FPMS is used to demonstrate programmable waveguides, amplifiers, and oscillators. A different implementation is demonstrated using LTCC and custom design CMOS chips. This one is a significantly smaller

implementation of the FPMS, and only waveguides of various width are shown. The loss and functionality of this circuit is not nearly as good as the FPMS fabricated using FR4. The higher loss can be remedied by using a custom LTCC stack-up and better switches. The frequency range of operation can be increased by locating the switches closer to the posts used in the unit cells. In addition, the size of the unit cells can be decreased further to increase performance.

Chapter 5

Discussion on the FPMS

A method of implementing programmable waveguides is presented. These waveguides exhibit low loss and extensive programmability as shown with two implementations of the FPMS. The material presented establishes the potential utility of the FPMS, however it still only scratches the surface. This chapter gives a brief qualitative discussion of various aspects of the FPMS. This ranges from increasing performance of the FPMS, briefly discussing limitations, to areas that require further investigation.

5.1 Extending Operating Frequency Range

The two examples of the FPMS shown previously both demonstrate limitations in operating frequency. The lower operating frequency is normally limited by the physical size of the device and this is a miniaturization issue. The upper frequency of operation is limited by the self-resonant frequency of the unit cell. The best method to increase the self-resonance frequency is to shrink the size of the unit cell. Shrinking the unit cell size reduces the capacitance associated with it and thus increase the self-resonant frequency. The unit cell sizes used in this dissertation are not close to

the limits of technologies readily available and thus the cell size can be miniaturized further. As seen in the LTCC implementation, the upper frequency is reduced by the metal lines required to connect to the CMOS chips. In any implementation, the package parasitics or additional metallization required to connect components has negative effects. This must be minimized or eliminated to increase upper frequency of operation. This can be done on LTCC by using an LTCC process that has MEMS switches.

5.2 Miniaturization

Miniaturization is a common theme in electronics and is very important. The programmable waveguides implemented in this work are based on microwave design concepts and are thus limited in size by the guided wavelength. The guided wavelength can be reduced by increasing the permeability and permittivity. However, this only reduces the wavelength by a factor of $1/\sqrt{\epsilon\mu}$. Miniaturization results in an increase of the resonant frequency of the unit cell. Using a single pole multi-throw switch allows for various capacitances to be selected, thus allowing for a large range of frequencies to be supported. At low frequencies a large capacitance can be selected to synthesize a very large effective dielectric constant, while at high frequencies a small effective dielectric constant can be realized with the switch off resulting in the only additional capacitance being the switch parasitics.

5.3 Unit Cell Shape

The implementations of the FPMS shown in this thesis utilize square unit cells. Future implementations can use hexagonal shaped unit cells. Using a different shape

allows for better control of electromagnetic fields in the programmable substrate.

5.4 Power Consumption

The FPMS power consumption is dictated by the switching technology used. The majority of switches only require power during switching and usually consume negligible power when in the on- or off-state. This means that other than during programming of the FPMS, the device consumes a negligible amount of power.

5.5 Power Handling

SIW circuits are demonstrated that can handle power levels in the range of 100W, and are more vulnerable to heating effects than dielectric breakdown [142]. This means that the components used to realize programmable behavior in the FPMS are the main limiting factor for power handling. Shrinking the FPMS unit cells results in increased power handling capabilities because the EM power is spread over more components, thus decreasing peak power or voltage on any one component.

5.6 Transitions

Transitions from microstrip or coplanar waveguides are important to implement practical circuits. Many of the transitions demonstrated for SIW circuits work well for the FPMS. However, many of the more compact transitions require careful design to adapt them for use in the FPMS. This is seen with the LTCC CPW transition demonstrated in the previous chapter. The center conductor of the CPW line is not capacitively loaded by the posts connected to the CMOS chips, which results in the

effective dielectric constant for the CPW being too small. This results in a poor return loss. It is important to develop good transitions for a variety of circumstances.

5.7 Traditional Slab Waveguide

It is possible some applications would benefit from the traditional slab waveguide that consists of a high dielectric constant core sandwiched between lower dielectric constant sides. It should be relatively easy to design an FPMS to realize this type of waveguide. The lower dielectric constant material sides may yield higher Q resonators and thus lower loss filters. The negative permittivity material can be used to isolate waveguides from each other or other circuits.

5.8 Antenna

It should be possible to adapt some SIW antennas to the FPMS. One such antenna is the half-mode SIW leaky-wave antenna [143]. The frequency of operation for this antenna is heavily dependent on the width of the leaky-wave section and this is easily changed with the FPMS. It would be desirable to realize antennas with programmable radiation patterns in addition to frequency reconfigurability.

5.9 Adapting SIW Circuits

Since the waveguides realized within the FPMS are very similar to SIW circuits it should be possible to easily implement SIW based circuits to the FPMS. This is advantageous because there are a significant amount of SIW circuits reported in the literature, which includes full SIW transceivers and other SIW based devices.

5.10 Transceiver

Most of the components required for a basic transceiver are presented. Demonstration of a full transceiver is a very important step and would help identify areas that need immediate attention.

5.11 Graceful Degradation

In any circuit, components will fail or performance will degrade over time. The FPMS can easily mitigate this problem by reprogramming it to adapt to a change in performance of components, or to route signals around failed components.

5.12 Self-test

Pre-designed and tested circuits should function fine on mass produced devices, but there may be variances and age will cause deterioration of performance. These issues can be mitigated with built-in self test circuitry. A full VNA implemented within the FPMS does not make sense. A method needs to be devised to allow the FPMS to test circuits implemented in it and optimize the performance of the circuits.

5.13 Hardware Description Language

Circuit design using the FPMS can proceed in a similar way to that currently done for microwave design. However, to make the device more accessible a hardware description language to program the FPMS would be highly advantageous. The goal of the language would be to allow designers to implement microwave circuits with similar ease to that done with digital circuits using verilog (i.e. provide a high level

of abstraction).

5.14 Realizing Unique Circuits

A programmable platform allows for circuits to be rapidly implemented and tested. This allows for genetic algorithms and other optimization algorithms to be ran using the FPMS to find circuits with unique performance. One would only need to set s-parameters that are desired and the FPMS could be used to find a circuit that yields the desired s-parameters. It is possible that complex circuits can be realized in simple and non-obvious ways using a method such as this. This kind of design is applied for antenna design and to realize unexpected analog circuit behavior on FPGAs [144]. Some circuits are not practical to design and fabricate because they are too sensitive to the inevitable fabrication variations. These circuits can be implemented in the FPMS because it can be easily fine tuned to operate as desired.

5.15 Programming Speed

The programming speed of FPMS based devices is only limited by the switching component. High switching speed can be achieved using semiconductor based devices, while devices that do not require as fast of switching speeds can use PCM or MEMS switches.

5.16 Special Circuitry

It is seen with FPGAs that there are instances where it makes sense to include certain digital circuits rather than implement everything in the programmable fabric. It is the same for FPMS. Certain components like high-Q resonators are difficult to

realize in the FPMS fabric and some applications may justify special circuitry. The FPMS fabric allows for signals to be easily routed to the additional components that cannot be realized in the programmable fabric.

5.17 Future Implementations

The state-of-the-art switches presented in Chapter 2 show excellent performance. It would be good to utilize some of these switching technologies to implement high performance FPMSs. MEMS technologies are demonstrated to be compatible with LTCC. Realizing a FPMS in LTCC with MEMS allows for a very high performance device that can cover a broad range of frequencies. Other switch technologies like PCM and semiconductor switches may be better suited to PCB implementations. Recently, there are demonstrations of mass-production assembly lines embedding components within the PCB in addition to surface mounted components, thus significantly increasing component counts [145]. A PCB implementation that makes use of the embedded PCB components and a multi-layer microwave PCB (i.e. Rogers 4350 substrates) can achieve high performance covering a broad frequency range and have a relatively small footprint. Liquid metal can be used to implement an FPMS as well. Implementations using liquid metals requires the most work, but they could yield some very promising capabilities not easily possible with other methods.

5.18 Conclusions

The work presented in this thesis is the foundation required to realize an FPMS and demonstrates two prototype circuits. There remains a large amount of work to fully develop the ideas presented in this dissertation. This chapter briefly discusses areas

that require further development.

Chapter 6

Conclusion

6.1 Summary

The purpose of this dissertation is to realize a method that yields low loss programmable microwave waveguides. Previous methods presented to realize programmable waveguides result in a significance amount of loss that makes the programmable waveguide unusable. The only programmable microwave circuits demonstrated or envisioned to date utilize basic microwave circuit building blocks (i.e. amplifiers, couplers, oscillators, etc..). These building blocks are sometimes reconfigurable themselves. A fabric, that consists of series switches, can dynamically connect these building blocks together to realize various transceivers. The success of these approaches is based on minimizing the number of switches required to realize a transceiver chain and using switches with very low insertion loss and good isolation. This is more of a 'coarse grained' approach, while the programmable waveguide approach is a 'fine grained' approach. The 'fine grained' approach yields a more versatile circuit and functional microwave circuits.

The programmable waveguides are realized using a device presented in this thesis called a field programmable microwave substrate. The substrate is composed

of a significant number of programmable unit cells. These unit cells can be programmed with a range of dielectric constants or be switched to a negative dielectric constant. It is determined that more and smaller unit cells yield better performance. The FPMS is programmed to realize slab waveguides with a positive dielectric constant core and negative dielectric constant sidewalls. These waveguides are used in conjunction with a transistor to realize a multitude of circuits such as waveguides, amplifiers, and oscillators.

6.2 Limitations

The FPMS is intended to realize programmable microwave waveguides and these can only practically be reduced in size to a certain point by increasing the effective permeability and permittivity. The programmable microwave circuits demonstrated in this work are able to mitigate the effects of lossy components significantly, however they can never function as well as application specific microwave circuits. There are power handling limitations to this type of device and this is likely in the range of 100W.

6.3 Implications

The FPMS presented in this dissertation provides a method to realize a general purpose microwave platform. It is envisioned that this kind of platform can replace some expensive and risky application specific microwave circuits in a similar way as the FPGA did with digital ASICs. In addition, it is expected that the FPMS can be used in applications that require a single platform to quickly switch between frequency bands and waveforms. It is expected that due to the high level of programmability

this device provides, people will devise completely unexpected uses for it.

6.4 Contributions

This work produced two contributions to the literature. The first contribution is indirectly related to this dissertation and is entitled "Modeling Substrate Integrated Waveguide Structures Using Effective Material Properties" [131]. This was presented at a conference (NEMO 2015) and demonstrates that substrate integrated waveguides can be interpreted as slab waveguides. In addition, the paper shows a method to significantly speed up simulations of SIW circuits by replacing the post wall of the waveguides with a homogeneous negative dielectric constant material.

The main contribution of this dissertation is a paper entitled "The Field Programmable Microwave Substrate". This paper was submitted to Microwave Theory and Techniques and is currently under review. This paper describes the important results presented within this dissertation.

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