Compact, Lightweight and Power Efficient Voltage Tunable Multiferroic RF/Microwave Components

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Abstract

In this dissertation, the knowledge of the partially magnetized ferrite concept and magnetoelectric coupling is discussed. The partially magnetized ferrite is able to operate in a very low permeability range less than unity with a low bias magnetic field, which is usually less than 100 Oe. The rapid fraction variation of the permeability will result in a large tunability in device application as the frequency response is closely related to the change of permeability. The magnetoelectric coupling of the ferrite can produce an effective magnetic field inside the material which gives rise to the anisotropy tuning with the absence of the external magnetic bias. However, this effective magnetic field is usually at the level of several 10’s Oe which is fairly weak in getting large frequency domain tuning of the device. Thus, a combination of both concepts is presented throughout this dissertation, which gives a great opportunity in achieving voltage tunable microwave devices with large tunability.

A method of measuring the complex permeability using a CPW and a network analyzer is presented. The permeability spectra under varied magnetic field is discussed. The measured permeability spectra show a negative region which prohibits the wave propagation. Therefore, the energy of the RF source dissipated in the material, and the attenuation depends on the magnitude of the negative value. In
addition, a resonator was fabricated on the ferrite substrate. The device showed an absorption band gap under a magnetic field from 200 Oe to 600 Oe, which was in good agreement with the measured permeability spectra.

A planar compact bandpass filter at C-band on a partially magnetized YIG substrate was demonstrated with a large tunability of 380 MHz (6.1%), a low insertion loss of 1.1 dB to 1.25 dB under low magnetic field of 0 to 100 Oe. The bandpass filter on unsaturated ferrite substrate also showed $IP_{1dB}$ up to 18 dBm and over 30 dBm in saturated state. A Ku-band multiferroic tunable bandpass filter on nikel spinel with an electrical tunability of 270 MHz (2.1%) is presented. The experiment discussed the usage of the spinel’s high magnetostriction in microwave devices. The frequency tunability is closely related to the ferrite permeability operating region. In addition, the measured central frequency under varied electric field showed a “butterfly” behavior due to the magnetic hysteresis of the nickel ferrite substrate in response to the PMN-PT single crystal.

A planar compact phase shifter at C-band on partially magnetized YIG substrate was demonstrated with a large differential phase shift of 350° at 6 GHz, a low insertion loss of 3.6 dB to 5.1 dB under a low magnetic field of 0 to 100 Oe. The phase shift exhibited a maximum of 72 degree/decibel loss at 6.55 GHz. Electrically
tunable multiferroic phase shifters presented a differential phase shift up to 94° with 1-layer PMN-PT/ferrite structure, and over 115° on a stacked 2-layer PMN-PT/ferrite structure operating in bending modes. The result is improved by 20%. The device is able to operate without any magnetic bias and this leads to devices that are more compact and power efficient.
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Chapter 1: Introduction

Modern ultra wideband communication systems, radars, and metrology systems all need reconfigurable subsystems that are compact, low loss, small form factor, and power efficient [1]. Ferrite has been applied in RF/microwave devices for more than half a century due to the low-loss, high power, high resolution, and high reliability.

For the application of ferrites in tunable devices, one usually takes the advantage of magnetic property influence of the wave propagating characteristic. A simplest example is the phase shift of a ferrite-loaded transmission line that can be controlled by the magnetization of the ferrite ceramics. [2] Similarly, tunable ferrite bandpass filters [3-5], phase shifters [6-8], isolators [9], and circulators [10] have drawn attention to many researchers. Typically, the study of the magnetic properties of magnetic materials or ferrites would allow one to have a better understanding of the wave propagation in magnetic media and a thorough prediction of the performances of the microwave devices.

Conventionally, ferrites and magnetic materials are tuned by magnetic field. In the past 10 years, multiferroics consisting of multiple order parameters and cross-coupling between the orders provide an alternative to tune the magnetic properties or anisotropies with electric fields [11], which has been of great interests in microwave tunable devices [12, 13]. This dissertation theoretically and experimentally focuses on the magnetic
properties of the ferrite materials, wave propagation in magnetic media, and multiferroics application in microwave devices.

1.1. Background

The essential concept throughout this dissertation is based on Maxwell’s equation which gives a general form of electromagnetic (EM) wave propagation in a medium [14]

\[
\begin{align*}
\text{Gauss’s law} & \quad \nabla \cdot \mathbf{E} = \rho/\varepsilon \\
\text{Gauss’s law of magnetism} & \quad \nabla \cdot \mathbf{B} = 0 \\
\text{Faraday’s law} & \quad \nabla \times \mathbf{E} = -\frac{\partial \mathbf{B}}{\partial t} \\
\text{Ampere’s law} & \quad \nabla \times \mathbf{B} = \mu \mathbf{J} + \mu \varepsilon \frac{\partial \mathbf{E}}{\partial t}
\end{align*}
\]

where

- \(\rho\) is the charge density (\textit{coulombs} \(\cdot\) \textit{m}^{-3})

- \(\mathbf{B}\) is the magnetic flux density (\textit{weber} \(\cdot\) \textit{m}^{-2})

- \(\mathbf{J}\) is the current density (\textit{amperes} \(\cdot\) \textit{m}^{-2})

- \(\mu\) and \(\varepsilon\) are permeability (\textit{henris} \(\cdot\) \textit{m}^{-1}) and permittivity (\textit{farads} \(\cdot\) \textit{m}^{-1})
In real media, the permeability $\mu$ is a tensor due to the anisotropic and dispersive characteristics of the medium, in particular, yields

$$\mu = \begin{bmatrix}
\mu_{xx} & \mu_{xy} & \mu_{xz} \\
\mu_{yx} & \mu_{yy} & \mu_{yx} \\
\mu_{zx} & \mu_{zy} & \mu_{zz}
\end{bmatrix}$$ (1.2)

For an anisotropic, inhomogeneous, and dispersive medium, the elements in the tensor are complex, and have frequency and spatial dependence. In this dissertation, we will focus on the tensor nature of the permeability, and discuss the interaction between the magnetic materials and the EM wave propagation.

1.2. Microwave ferrites

Low loss ferromagnetic materials, or ferrites are widely used in passive RF/microwave components. The magnetic resonances of the components can be tuned over a wide frequency range using magnetic fields. There are basically three types of ferrites: hexaferrites, garnets, and spinels.

Hexaferrites have a hexagonal crystal structure and with a large saturation magnetization ($4\pi M_s$) and a large magnetocrystalline uniaxial anisotropy ($H_A$), and will bias the ferrite to high frequencies, such as Ku band. In particular, M-type (BaFe$_{12}$O$_{19}$) ferrites has an out of plane easy axis of magnetization which led to many circulator designs [2].
Garnets, as a class of very low loss material, have been applied in RF/microwave devices for many years. The first study of the cubic crystal structure of garnets is due to Menzer. Bertaut and Forrat prepared the most well-known garnet yttrium iron garnet (Y₃Fe₅O₁₂, or YIG). The YIG material exhibits very low loss at high frequencies. The ferromagnetic resonant (FMR) linewidth (ΔH) of a typical YIG is measured to be less than 1 Oe single crystals, and 25 Oe for polycrystallines. The $4\pi M_s$ can be varied from 1200 Gauss to more than 2000 Gauss with different doping composites. Therefore, the materials show great merit in military or commercial communication systems.

Although spinels exhibit a large initial permeability, the value drops very quickly with the increase of frequency. In addition, the spinels are known as high relaxation loss materials with a typical ferromagnetic loss (ΔH) over 150 Oe. The material is not comparable with garnets in microwave applications. However, the introducing of magnetoelectric (ME) coupling recently through electric fields is more attractive than magnetic control of magnetism which gives rise to the study on spinels. The saturation magnetostriiction ($\lambda_s$) can be as high as several ten’s ppm which show great potential of achieving voltage control of magnetism. [15] For example, Nickel spinel has a $\lambda_s$ of 25 ppm compared to YIG garnet of -1~2 ppm.
1.3. Microwave and magnetic properties

1.3.1. Permeability tensor

Consider which a magnetic dipole merged in a constant magnetic field along the z-axis. At microwave frequencies, the magnetization of a system of a magnetically aligned spins or the magnetic moment per unit volume $M$ follows the equation of motion in vector form [16]

$$\frac{dM}{dt} = -\gamma (M \times H) \quad (1.3)$$

where $\gamma = 2.8MHz/Oe$ is the gyromagnetic factor, the symbol $H$ and $M$ represents the vector sum of all the magnetic field and magnetization. The total magnetic field and magnetization is given by

$$H_t = \hat{z}H_0 + he^{j\omega t} \quad (1.4a)$$

$$M_t = \hat{z}M_0 + me^{j\omega t} \quad (1.4b)$$

where $H_0$ is the applied magnetic field, $M_0$ is the dc magnetization, and the rf term $he^{j\omega t}$ contains information of the amplitude $h$ and the frequency $\omega$ of the microwave field. $m$ is the magnetization induced by rf field $h$. By substituting eq. (1.4) into eq (1.3) yields

$$\frac{dm}{dt} = -\gamma (\hat{z}H_0 + h) \times (\hat{z}M_0 + m) \quad (1.5)$$

In the case where $H_0$ is sufficiently large that the magnetic dipole are all aligned along the magnetic field, $H_0 \gg h$, and $M_0 \gg m$, one can obtain eq (1.5) in form of
\[
\begin{align*}
\frac{dm_x}{dt} &= -\omega_0 m_y + \omega_m h_y \\
\frac{dm_y}{dt} &= \omega_0 m_x - \omega_m h_x \\
\frac{dm_z}{dt} &= 0
\end{align*}
\] (1.6)

where \(\omega_0 = \gamma H_0\), and \(\omega_m = \gamma M_0\). Take derivative over time on both sides of eq (1.6) yields

\[
\begin{align*}
(\omega_0^2 - \omega^2)m_x &= \omega_0 \omega_m h_x + j\omega \omega_m h_y \\
(\omega_0^2 - \omega^2)m_y &= -j\omega \omega_m h_x + \omega_0 \omega_m h_y
\end{align*}
\] (1.7a)

or

\[
m = \begin{bmatrix}
\chi_{xx} & \chi_{xy} & 0 \\
\chi_{yx} & \chi_{yy} & 0 \\
0 & 0 & 0
\end{bmatrix} h
\] (1.7b)

where \(\chi_{xx} = \chi_{yy} = \frac{\omega_0 \omega_m}{\omega_0^2 - \omega^2}\), \(\chi_{yx} = -\chi_{xy} = \frac{j\omega \omega_m}{\omega_0^2 - \omega^2}\). Knowing the relation \(b = \mu_0(m + h)\),

the tensor permeability in saturated case then can be written as

\[
\mu = \mu_0 \begin{bmatrix}
1 + \chi_{xx} & 1 + \chi_{xy} & 0 \\
1 + \chi_{yx} & 1 + \chi_{yy} & 0 \\
0 & 0 & 1
\end{bmatrix} = \mu_0 \begin{bmatrix}
\mu & -j\kappa & 0 \\
j\kappa & \mu & 0 \\
0 & 0 & 1
\end{bmatrix}
\] (1.8)

where

\[
\mu = 1 + \chi_{xx} \quad \quad j\kappa = -\chi_{xy}
\]

For the case \(H_0\) is not sufficiently large, the materials is unsaturated, the elements in tensor permeability follows a more complicated expression. This will be discussed in chapter 3.
1.3.2. Demagnetizing field

Magnetic materials show an intrinsic field to oppose the external magnetic field. It is produced by the surface magnetic charge on the interface between the magnetic and nonmagnetic material. Consider a finite ferrite thin plate with z-axis normal to the board (Fig. 1.1a). The magnetic field is applied either along z-axis or in-plane. Assume all the magnetic moment is aligned along the magnetization direction. If the magnetic field is applied along the z-axis, Gauss’s law gives

\[ B_o = B_i, \quad (1.9a) \]

\[ B_o = H_o, \quad (4\pi M = 0 \text{ in air}) \quad (1.9b) \]

\[ B_i = 4\pi M + H_i \quad (1.9c) \]

where \( H_i \) is the internal field, \( H_o \) is the applied external magnetic bias, and \( M \) is the magnetization along the z-axis. Eq (1.9a-c) yields

\[ H_i = H_o - 4\pi M \quad (1.10) \]

From eq (1.10) one obtains the demagnetizing field of a ferrite plate with z-axis normal to the board face is \( 4\pi M \), and the z-axis demagnetizing factor is \( N_z = 4\pi \).

Similarly, for a tangential external magnetic bias (fig. 1.1b), the magnetic field needs to be continuous at the interface of the magnetic and nonmagnetic surface, giving
Therefore, one can conclude the x- and y-axis demagnetizing factor for a thin plate is $N_x = N_y = 0$. 

\[ H_i = H_0 \]

(a)
1.3.3. Remanant magnetization

After the ferrite is magnetized to saturation, the magnetization will relax to a remanant magnetization ($M_r$) with the absence of the external magnetic bias. For YIG and Nickel spinel, the $M_r$ is quite small. (Fig. 1.2) However, for M-type strontium hexaferrite with an out of plane easy magnetization, the $M_r$ may be as large as 3500 Gauss, hence is a good candidate for self-biased junction circulators [2]. It has also been used in many latching wire devices [17].
Fig. 1.2 Remanant magnetization of YIG and Nickel spinel.

1.4. Multiferroics

Multiferroics have been formally defined as materials that exhibit more than one primary ferroic order parameter simultaneously. Fig 1.3 shows the relationship between multiferroic and magnetoelectric materials [18]. The red area represents materials that are multiferroic. As independent orders, magnetization and polarization in multiferroics is able to realize four logic states and enhance the functionality in multiferroic devices. But in practice, besides both independent orders, the cross-coupling between them would be more attractive, since it reveals a dual-tunable capability that is a dielectric polarization variation in response to an applied magnetic field, or an induced magnetization from an
external electric field. In this dissertation, we focus on the latter in microwave device application, which is so-called the converse magneto-electric (ME) coupling. Generally, ME coupling can exist in any magnetic and electrical orders and may arise directly between them.

Fig. 1.3 The relationship between multiferroic and magnetoelectric materials. – The relationship between multiferroic and magnetoelectric materials [18].

However, it also enables achievement of interfacial multiferroic coupling through a mechanical channel in heterostructures consisting of a magnetoelastic and a piezoelectric component [11]. The strain mediated indirect ME coupling usually occurred in a certain kind of material which is composed of separate magnetic and electrical phases.
1.5. Dissertation overview

This dissertation focuses on the advantages of partially magnetized ferrite usage in magnetoelectric coupling (ME) devices, the combination of these two concepts is proposed. The magnetic permeability of partially and fully magnetized ferrite is theoretically and experimentally studied. The EM wave propagation in ferrite media will be discussed. In addition, voltage tunable multiferroic microwave devices such as phase shifters and bandpass filters are implemented.

In chapter 2, I will briefly introduce the numerical modeling tools, fabrication tools, and the measurement setups that are used throughout this dissertation. First, HFSS software is introduced as a pre-experimental tool. Next, some key fabrication tool such as spin coater and the mask aligner which are used to do the photolithography process are described. Finally, the setups that are used for characterizing the fabricated devices, such as vector network analyzer are discussed.

In chapter 3, we will provide a theoretical overview of the permeability tensor in the states before saturation. The Schloemann’s model and the Naito’s model will be introduced. Green’s experimental results will also be discussed to verify the models’ consistency. Furthermore, magnetoelectric coupling theory is also given for an advanced
application of the partially magnetized ferrites. The concept of combining the two theories is proposed as the basis of this dissertation.

In chapter 4, the frequency and field response of the ferrite permeability in fully saturated states will be studied. First, I will give an introduction to the research state of art on the measurement of permeability techniques. Next, a simple but effective approach of measuring the permeability will be described, and the experiment result of the permeability measurement is presented. Then, a resonator on ferrite substrate is presented, and the test frequency response is in agreement with the permeability measurement. In addition, a prediction of the wave propagation prohibiting band gap is proved by the negative region from the permeability measurement.

In chapter 5, bandpass filters on partially magnetized ferrite concept is introduced. Besides, the state of art of voltage tunable multiferroic bandpass filters is discussed. To implement a voltage tunable multiferroic bandpass filter using partially magnetized concept, a low loss bandpass filter at C-band on YIG substrate as a startup is presented. The bandpass filter shows a tunability of over 6% with less than 100 Oe magnetic bias, the insertion loss is reported as low as 1.1 dB. In addition, the power handling capability of the devices is presented with an IP_{1dB} of 18 dBm. Nickel spinel ferrite with a high saturation magnetostriction value of ~ -33 ppm is used for multiferroic bandpass filter at
$K_U$–band. The bandpass filter shows an electric field tunability of 270 MHz through ME coupling under a 100 Oe magnetic bias.

In chapter 6, a planar compact phase shifter at C-band on partially magnetized YIG substrate was demonstrated with a large differential phase shift of 350° at 6 GHz, a low insertion loss of 3.6 dB to 5.1 dB under a low magnetic field of 0 to 100 Oe. The phase shift exhibited a maximum of 72 degree/decibel loss at 6.55 GHz. A differential phase shift up to 94° is obtained with the PMN-PT/ferrite heterostructure. Planar electrically tunable multiferroic phase shifter on a stacked 2-layer PMN-PT/ferrite structure was demonstrated with a large differential phase shift over 115° under an electric field of 11 kV/cm. The result is improved by 20% compared to 1-layer PMN-PT design. The device is able to operate without any magnetic bias leads to devices that are more compact and power efficient. The devices combining with multiferroic and partially magnetized ferrite concept is promising in achieving voltage tunable devices with large tunability.

Chapter 8 will be the conclusion.
1.6. References


Chapter 2 : Simulation, Fabrication and Measurement Setups

In this chapter, the numerical and experimental setups for the simulation, fabrication and measurement of the microwave devices are presented. The modeling software High Frequency Structural Simulator (HFSS) from ANSYS will be introduced. Some of the key fabrication equipment will be listed. In addition, the setups for the measurement of the devices will also be covered. [1]

2.1. High Frequency Structural Simulator (HFSS)

HFSS is a commercial finite element method solver for electromagnetic structures from ANSYS. It is one of several commercial tools used for antenna design, and the design of complex RF electronic circuit elements including filters, transmission lines, and packaging. Characterization of the model, such as field distribution, S-parameters, radiation pattern, phase, resonant frequency, etc. can also be obtained from the software.

When magnetic materials, e.g. ferrites, are involved in the design, the HFSS can only model the ferrite materials in fully saturated states, where the permeability is a tensor [2]

\[
\boldsymbol{\mu} = \begin{pmatrix}
\mu & -j\kappa & 0 \\
jk & \mu & 0 \\
0 & 0 & \mu_z 
\end{pmatrix}
\]  \hspace{1cm} (2.1)
\[ \mu = \mu_0 \left( 1 + \frac{\omega_0 \omega_m}{\omega_0^2 - \omega_m^2} \right), \kappa = -\mu_0 \frac{\omega \omega_m}{\omega_0^2 - \omega_m^2} \]

where \( \mu_z = 1 \) for saturated case. \( \omega_0 = \gamma H_0, H_0 \) is the dc magnetic bias; \( \omega_m = \gamma 4\pi M_s \), \( M_s \) is the saturation magnetization of the ferrite, and \( \omega \) is the signal angular frequency. When stronger magnetic field beyond saturation is applied, the magnetization remains as the value \( M_s \).

However, when the ferrite is operated in partially magnetized states, the actual magnetization \( M \) is a value between 0 and \( M_s \) depending on the magnitude of field applied. The diagonal element of in the tensor permeability follows a more complex expression. Therefore, it is hard to characterize the model with respect to the change of magnetic field in HFSS. In this dissertation, we use numerical combining with software modeling as the simulation.

### 2.2. Fabrication facilities

The devices described in this dissertation do not follow the procedure for fabrication on standard printed circuit boards (PCBs). Cleanroom and MEMS process are involved but with a much large scale and error tolerance. Here listed some of the key equipments that have been used in the experiments.
2.2.1. Physical Vapor Deposition (PVD) system

PVD is a variety of vacuum deposition methods used to deposit thin films by the condensation of a vaporized form of the desired film material onto various workpiece surfaces. The coating method involves purely physical processes such as high temperature vacuum evaporation with subsequent condensation, or plasma sputter bombardment rather than involving a chemical reaction at the surface to be coated as in chemical vapor deposition (CVD). PVD coatings are sometimes harder and more corrosion resistant than coatings applied by the electroplating process. Most coatings have high temperature and good impact strength, excellent abrasion resistance and are so durable that protective topcoats are almost never necessary.

Energetic ions sputter material off the target which diffuse through the plasma towards the substrate where it is deposited. There is no strong plasma glow around the cathode since it takes a certain distance for the plasma to be generated by electron avalanches started by a few secondary electrons from the sputtering process. (Fig. 2.1)
2.2.2. Laurell spinner for spin coating

Spin coating is a procedure used to apply uniform thin films to flat substrates. An excess amount of a solution is placed on the substrate, which is then rotated at high speed in order to spread the fluid by centrifugal force. Rotation is continued while the fluid spins off the edges of the substrate, until the desired thickness of the film is achieved. Spin coating is widely used in microfabrication photolithography process, to deposit layers of photoresist.
2.2.3. Quintel 4000 Mask Aligner

Quintel 4000-6 mask aligner is used for UV exposure after the photoresist (PR) has been coated onto the substrate sample. After prebaking, the photoresist is exposed to a pattern of intense light. The exposure to light causes a chemical change that allows some of the photoresist to be removed by developer. The exposure process is illustrated in Fig. 2.3.
In addition, Inductively Coupled Plasma (ICP Plasma Therm 790) and Veeco Microetch Ion Mill are used for dry etch.

2.3. Measurement tools

2.3.1. Vector Network Analyzer (VNA)

The characterization of the microwave devices are mainly carried out by our VNA (Agilent PNA E8364A), 45 MHz to 50 GHz. The equipment can produce a power level up to 30 dBm, which will be used for power handling capability measurement in this dissertation.
Fig. 2.4 Network analyzer Agilent PNA E8364A.
2.3.2. Lock-in Amplifier

The Model SR830 DSP Lock-In Amplifier (Stanford research systems) connecting with a High-Voltage Power Amplifier (gain 1000V/V) produces a voltage of $\pm 600 \text{ V}$.

Fig. 2.5 Model SR830 DSP Lock-In Amplifier from Stanford research systems.

2.4. References

1. HFSS V10 user guide.

Chapter 3: Theory of Partially Magnetized Ferrite and ME Coupling

3.1. Partially magnetized ferrites

In many of the application of ferrites in microwave devices such as phase shifters, filters, switches, and rotators the magnetic material is only partially magnetized. The unsaturated ferrite exhibits a low permeability value less than unity and a significant variation is obtained with a modest field. As in Fig. 3.1, a closed path magnetic field of 10~20 Oe is able to magnetize the ferrite toroid (gadolinium yttrium iron garnet) to saturation [1].
In device point of view, the fractional variation of the permeability will result in a large change in frequency domain, since either the phase shift of a phase shifter or the resonant frequency of a bandpass filter relies on the factor $\sqrt{\mu}$ or $1/\sqrt{\mu}$. Therefore, understanding of the partially magnetized ferrite mechanism will provide a helpful guidance to design the microwave devices.

### 3.1.1. Schloemann’s model

Assuming there are only two types of domains in this model, which are aligned either parallel (“up domains”) or anti-parallel (“down domains”) to the z-axis. This provides a cylindrical domain configuration with arbitrary cross sections in the x-y plane. The dc magnetizing field is applied in z-direction, thus the dc magnetization is either parallel to the z – direction (not as up domains “u”), or anti-parallel to the -z – direction (not as down domains “d”) as shown in Fig. 3.2. In the plot, two or more “up domains” or “down domains” are adjacent, then the domains with the same type can be combined into one domain.
Fig. 3.2 Domains with magnetization parallel (“u”) or anti-parallel (“d”) to z-direction.

The local permeability is the same throughout each type, but differs from one type of domain to the other. Then the $x$ and $y$ components of the local permeability tensors for the two types of domains can be expressed by

$$
\mathbf{\mu}_u = \begin{pmatrix} \mu_u & -i\kappa_u \\ i\kappa_u & \mu_u \end{pmatrix}, \quad \mathbf{\mu}_d = \begin{pmatrix} \mu_d & -i\kappa_d \\ i\kappa_d & \mu_d \end{pmatrix}
$$

(3.1)

from equations of motion discussed in chapter 1, one can obtain
\[ \mu_{u,d} = 1 + \frac{4\pi M_0 H_{u,d}}{H_{u,d}^2 -(\omega/\gamma)^2} \] (3.2a)

\[ \kappa_{u,d} = -\frac{4\pi M_0 \omega/\gamma}{H_{u,d}^2 -(\omega/\gamma)^2} \] (3.2b)

where the \( H_{u,d} \) is the dc magnetic field strength in \( z \) – direction. Under the condition described in [2], it is permissible to set \( \mu_u = \mu_d, \kappa_u = \kappa_d \). If the loss is not taken into account one can obtain

\[ \mu = 1, \text{ and } \kappa = \omega_m/\omega \]

where \( \omega_m = \gamma 4\pi M_0 \).

When an rf magnetic field \( h(r) \) is associated with some arbitrary distribution of rf magnetization \( m(r) \), the \( h(r) \) is a gradient of a potential \( h(r) = \nabla \Psi(r) \). By knowing the divergence of \( b = \mu \cdot h \) vanishes, \( \Psi(r) \) then satisfies

\[
\left[ \mu \left( \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} \right) + \frac{\partial^2}{\partial z^2} \right] \Psi = 0
\] (3.3)

In this configuration it is \( z \) independent, thus eq (3.3) yields

\[
\left( \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} \right) \Psi = 0
\] (3.4)

The boundary condition of an adjacent “up domain” and “down domain” requires the normal component of \( b \) and the tangential component of \( h \) must continuous.
where “n” and “s” represents the normal and tangential component at the interface. Eq (3.4) should also satisfy the boundary condition at the interface between the material and the air.

The effective circularly permeability $\mu_{+eff}$ and $\mu_{-eff}$ is given in response of the sense of positively and negative rotation of rf magnetic field, respectively, which have the form

$$\mu_{\pm eff} = \mu_{eff} \mp \kappa_{eff}$$

(3.6)

A simplified cylindrical domain structure is used for the solution the differential eq (3.5). The domain structures observed in x-y plane are concentric circles with “up domain” and “down domain” periodically aligned with z – axis, as shown in Fig. 3.3. The radii of the domains are $r_n (n=1, 2, 3, \cdots)$. Therefore, the solution follows the forms of

$$\psi = r^m e^{-im\phi}, r^{-m}e^{-im\phi}$$

(3.7)

where m is positive integers. Specifically, the potential of nth domain is expressed as

$$\psi_n = (A_nr + B_nr^{-1})e^{-i\phi}$$

(3.8)
At the core domain, $B_n = 0$. Outside the sample, $A_{N+1} = h_o$ to satisfy the $r \rightarrow \infty$ condition, where $h_o$ is the amplitude of rotated rf magnetic field.

![Diagram](image.png)

**Fig. 3.3** A simplified domain configuration with concentric cylindrical structure.

A complex mathematical calculation leads to:

$$
\mu_{eff}^2 - \kappa_{eff}^2 = \mu^2 - \kappa^2
$$

(3.9)

This gives a similar result as Rado’s model [3]. The right hand side of eq (3.9) is independent to the domain configuration. Therefore, the minimum $\mu_{eff}$ regardless of domain configuration is obtained when $\kappa_{eff}$ is zero.
where $H_a$ is assumed as the only anisotropy field that exists inside the materials. A detailed calculation may be found in [2]. The discussion so far the permeability correspond to x and y – axis, for a three dimensional case, the effective permeability corresponds to z – axis is equaled to unity which will contribute 1/3 of the effective permeability. Therefore, the minimum effective permeability in demagnetized state yields

$$
\mu_{eff \ min} = \mu_{demag} = \frac{2}{3} \left[ \frac{\left(\omega/\gamma\right)^2 - (H_a + 4\pi M_0)^2}{(\omega/\gamma)^2 - H_a^2} \right]^{\frac{1}{2}} + \frac{1}{3}
$$

(3.11)

It is seen from eq (3.11), that when $\omega > \gamma(H_a + 4\pi M_0)$, the $\mu_{demag}$ is a real number; while for $\gamma H_a < \omega < \gamma(H_a + 4\pi M_0)$, $\mu_{demag}$ has an imaginary part. This indicates that the material will introduce addition loss. Assuming $H_a$ is not comparable to $\gamma 4\pi M_0$, the term leaves a so-called “low field loss” condition, i.e. $\gamma 4\pi M_0 / \omega = \omega_m / \omega$ is close to or great than unity. The calculated curve in terms of $\omega_m / \omega$ fits the measurement result for different types of ferrites [4], as shown in Fig 3.4.
Fig. 3.4 Measured real part of permeability of ferrite in demagnetized state. $-\mu'_0$

versus $\gamma 4\pi M/\omega$. [4]
$\mu_0^{11} \times 10^3$ $\gamma 4\pi M_s/\omega$
Fig. 3.5 $\mu'_0$ versus $\gamma 4\pi M_0/\omega$ for various Tran Tech garnets. $-\mu''_0$ versus $\gamma 4\pi M_0/\omega$ for various Tran Tech garnets. [5]

However, the investigation to eq (3.11) implies that for a smaller $\omega_m/\omega$, a lower $\mu_{demag}$ is obtained, and the tunability of microwave device will take its advantage. On the other hand, with a smaller $\omega_m/\omega$, the material is more close to the “low field loss” region which will degrade the insertion loss of the device. More specifically for $\omega_m/\omega < 0.7$, the imaginary part is given by [5]

$$
\mu''_0 = \frac{\gamma \Delta H_{eff} \gamma 4\pi M_0}{2 \omega^2}
$$

(3.12)

3.1.2. Naito’s model

The Schlamann’s model is not applicable when dc magnetic field is applied over the sample. On the basis of Schlamann’s model, the Naito’s model of ferrite in any partially magnetized state will be discussed in this section.

In this model, the domains still follows the configuration in Fig. 3.2 where the magnetization orientation of each domain is either parallel or anti-parallel to the $+z$ – direction. Define the volume of parallel and anti-parallel domains are $V_p$ and $V_a$, and $V_p + V_a = V$, where $V$ is the total volume. Similarly, $\mu_+ = \mu - \kappa$ and $\mu_- = \mu + \kappa$ are
permeabilities in response to the positive and negative rotation of the rf magnetic field. Here, if a frequency dependent loss term $\alpha$ is introduced, one can write $\mu_+, \mu_-$, and $\mu$ as

$$\mu_+ = \mu - \kappa = 1 + \frac{\omega_m}{\omega + \omega_e + j \omega \alpha} \quad (3.13)$$

$$\mu_- = \mu + \kappa = 1 + \frac{\omega_m}{\omega + \omega_e + j \omega \alpha} \quad (3.14)$$

$$\mu = 1 + \frac{\omega_m(\omega_e + j \omega \alpha)}{-\omega^2 + (\omega_e + j \omega \alpha)^2} \quad (3.15)$$

where $\omega_e$ is a term related to the effective dc magnetic field. $\mu$ is the diagonal element in the tensor permeability of “parallel” and “anti-parallel” case. If the rf magnetic field is linear and in the $x - y$ plane, the diagonal element $\mu_\perp$ is expressed based on an arbitrary but reasonable assumption as [6]

$$\mu_\perp = \sqrt{\mu_+ \mu_-} \left( \frac{2V_p}{V} \right) \left( \frac{2V_a}{V} \right) + \mu \left\{ 1 - \left( \frac{2V_p}{V} \right) \left( \frac{2V_a}{V} \right) \right\} \quad (3.16)$$

When there is no dc magnetic field, the magnetization equals to zero, thus $V_p = V_a = \frac{1}{2} V$ yields

$$\mu_\perp = \sqrt{\mu_+ \mu_-} \quad (3.17)$$

Eq (3.17) is the case of eq (3.9) except the loss term. When small dc magnetic field is gradually applied along $+z$ – direction, a volume of $\delta V$ anti-parallel domains is turned $180^\circ$ into parallel domains. In this case, the difference between two types of domains is
It is reasonable to consider the magnetization \( M \) is proportional to \( 2\delta V/V \). We have \( 2\delta V/V = M/M_0 \). Eq (3.16) is then written as

\[
\mu_\perp = \sqrt{\mu_\parallel \mu_\perp} \left\{ 1 - \left( \frac{M}{M_0} \right)^2 \right\} + \mu_\perp \left( \frac{M}{M_0} \right)^2
\] (3.18)

Still, the \( z \) – component contributes \( 1/3 \) to the total effective permeability. Finally, the effective permeability follows

\[
\mu_{\text{eff}} = \frac{2}{3} \left\{ \sqrt{\mu_\parallel \mu_\perp} \left\{ 1 - \left( \frac{M}{M_0} \right)^2 \right\} + \mu_\perp \left( \frac{M}{M_0} \right)^2 \right\} + \frac{1}{3}
\] (3.19)

Compared to Schloemann’s formula, this expression not only induced the loss term, but also can be applied to any partially magnetized state, once the variables: the single frequency \( \omega \), the magnetization \( M \), and the dc magnetic field \( H_o \) are known.
Fig. 3.6 Theoretical curve versus measured scatters. – $\mu'$ versus $H_N$ on TTI-390 (MgMn ferrite with second phase) at 5.5 GHz, where $\mu'(H_i \to +0) = 0.95$, $\omega_e/\omega = 0.058$, and $b=80^\circ$ [6].

The “low field loss” is also discussed in this model. In discussion here only give the conclusion. Detail information may be found in [6]. The “low field loss” are absent under the condition that $\omega_m/\omega < 0.745$, and show a constant loss over the partially magnetized state. The “low field loss” increases when the ratio is greater than the threshold.

3.1.3. Summary

In this section, we discussed two models to predict the ferrite effective permeability in partially magnetized state. Schloemann’s model with a concentric cylindrical domain configuration gave the effective permeability in demagnetized state. Naito expended the model to a more general case where the dc magnetic field exists. Though the formula is empirically proposed, the theory can explain well to the Green’s measurement result on different types of ferrites. Besides, Gelin [7] presented a consistent model with taking into account the interaction between adjacent domains. Due to the dissertation structure, it will not be discussed.
3.2. Magnetoelastic (ME) coupling

For the two-phase systems, the physical properties are determined by the interaction between the constituents as well as by their individual properties. Some effects, which are already present in the constituents may be averaged or enhanced for the overall system. However, the ME effect is among the novel effects that arises from the product properties originating through the interaction between the two phases. The ME effect is implemented by the coupling between an electrostrictive material and magnetostrictive material via a good contact. [8]

Fig. 3.7 Pathways between electrical, magnetic, and elastic phase – Phase control in ferroics and multiferroics [9].
An applied magnetic field over the magnetostrictive phase will produce a strain on the material thus results in a mechanical deformation. The strain is then transferred to the electrostrictive phase and produces an electric polarization. This is procedure is so-called direct ME effect, as shown in Fig 3.7. Oppositely, an electric polarization on the piezoelectric will also induce a strain in the electrostrictive material, the mechanical deformation on the structure will produce a magnetization modulation in the magnetostrictive phase. This is so-called converse ME effect. The ME coefficient can be described by

\[
ME_{H-E} = \frac{\text{magnetic}}{\text{mechanical}} \times \frac{\text{mechanical}}{\text{electric}} \quad \cdots \quad \text{direct ME effect} \quad (3.20)
\]

\[
ME_{E-H} = \frac{\text{electric}}{\text{mechanical}} \times \frac{\text{mechanical}}{\text{magnetic}} \quad \cdots \quad \text{converse ME effect} \quad (3.21)
\]

Assuming a thin ferrite film is deposited on top of a piezoelectric material, as shown in Fig. 3.8. The electric field induced stress will lead to an in-plane stress on the ferrite material, which noted as \(\sigma_x\) and \(\sigma_y\) along \(x\) and \(y\) – axis. The ME energy is expressed as [10]

\[
F_{ME} = -\frac{3}{2}\lambda \sigma_x \sin^2\theta \cos^2\phi - \frac{3}{2}\lambda \sigma_y \sin^2\theta \sin^2\phi \quad (3.22)
\]

where \(\lambda\) is the magnetostriction of the ferrite material. The effective in plane anisotropy induced is calculated as the second order derivative upon \(\theta\) and \(\phi\) yields
\[
H_{\text{eff}, x} = \frac{3\lambda(\sigma_x - \sigma_y)}{M_0} \tag{3.23a}
\]

\[
H_{\text{eff}, y} = -\frac{3\lambda(\sigma_x - \sigma_y)}{M_0} \tag{3.23b}
\]

with

\[
\begin{pmatrix}
\sigma_x \\
\sigma_y
\end{pmatrix} = \frac{Y}{1-\nu^2} \begin{pmatrix}
1 & \nu \\
\nu & 1
\end{pmatrix} \begin{pmatrix}
d_{31} \\
d_{32}
\end{pmatrix} E \tag{3.24}
\]

where \(Y\) and \(\nu\) are the Young’s modulus and the Poisson’s ratio of the magnetic material. \(d_{31}\) and \(d_{32}\) are the piezoelectric coefficients, with a compressive stress in negative and tensile stress in positive.

---

**Fig. 3.8** A schematic of ME effect on a ferrite-piezoelectric multiferroic structure.

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### 3.3. Conclusion

In this chapter, two areas of knowledge are included. First, the theoretical model effective permeability of partially magnetized ferrite is presented. The Schlomann’s
model with concentric cylindrical domain configuration is used to calculate the ferrite
permeability in demagnetized state. Next, the Naito’s model is discussed as an expansion
of Schlomann’s permeability in any partially magnetized state with the existence of dc
magnetic field. The models agree with the measurement results by Green. Furthermore,
ME theory is given by calculating the energy of a two-phase multiferroic structure hence
to model the effective magnetic field in the magnetoelastic material.
3.4. References


Chapter 4: Ferrite Substrate Based Resonator

Magnetically control of the ferrite permeability is one of the approaches of implementing tunable devices. The study on the magnetic permeability is of importance in the design of ferrite devices. In this chapter, starting from a permeability measurement technique, the ferrite permeability spectrum under varied external magnetic bias is presented. The, a magnetically low loss tunable resonator at S-band is presented. The measured transmission coefficients show a band gap under a magnetic bias from 200 Oe to 600 Oe, where the permeability undergoes a negative value at the proposed frequencies. The test results are in good agreement with the permeability prediction.

4.1. Permeability prediction background

There are theoretical and experiment efforts on the prediction of the magnetic permeability tensor in terms of DC magnetizing field. The analytical expressions for $\mu$ and $\kappa$ in fully magnetized state were derived by Polder [1]; for unsaturated case, Gelin [2] gave a more precise model compared to the work had been done by Rado [3], Schlomann [4], and Naito [5]. Experimental prediction techniques on the permeability tensor by Krupka [6], Queffelec [7] and Ding [8] also give rise to a consistent evidence of those models. In devices point of view, the diagonal element in broadband frequency response is important for their applications. It will provide a better understanding of the wave propagation in magnetized ferrites and the performance of microwave devices.
Ding, et al. [8] reported a broad-band permeability measurement technique using a network analyzer and a coplanar waveguide (CPW). The main idea of the method is to derive the permeability value from the effect that the magnetic material brought into the nonmagnetic circuit. The schematic is shown in Fig. 4.1.

![Schematic of measurement setup in [8]](image)

Fig. 4.1 Schematic of measurement setup in [8].
A CPW is connected to the VNA, and two pairs of current driven magnetic coil is mounted to provide both longitudinal and in plane transverse magnetic field. After the circuit is well calibrated, the sample is placed on top of the CPW, which will result in perturbation of the non-magnetic circuit.

To measure the longitudinal biased permeability, first, a stronger bias magnetic field is applied along the y-axis to saturate the magnetic material. The transmission ($S_{21}$) and reflection ($S_{11}$) is recorded by a VNA after subtract the background noise, as $S_{21}$ and $S_{11}$. Since the S21 and S11 are measured when the ferrite is in saturated state, they the information regarding only the nonmagnetic properties of the circuit.

Next, a longitudinal magnetic bias is applied along the x-axis to saturate the sample. The magnetic bias is then reduced to a desired magnitude, and the material will then bring magnetic effect on the transmission and reflection. For transmission, notch band at certain frequency regions is observed. The s-parameters are then recorded as, $S'_{21}$ and $S'_{11}$. For simplicity, the permeability in a complex form is given by

$$
\mu = \frac{z_0 \left( \frac{1 + S'_{11} - S'_{21}}{1 - S'_{11}} \right)}{igt \mu_0 \omega} \left( \frac{1 + S'_{11} - S'_{21}}{1 - S'_{11}} \right)
$$

(4.1)

where $l$ is the length of the sample along the CPW, $t$ is the thickness, $g$ is a geometry factor which is used for data post-processing.
4.2. Magnetic properties measurements

4.2.1. Magnetic permeability broad-band measurement

In our experiments, a 10 mm coplanar waveguide is mounted onto a Cu fixture with two SMA connectors, the fixture is then connected to the two ports on the VNA. The ferrite samples are doped yttrium iron garnet (YIG) material prepared by MIT Lincoln laboratory (MIT-4). The initial permeability and permittivity is 10 and 13, respectively. The samples are cut into 10 mm × 5mm slabs with a thickness of 0.5mm and are used as the substrate of the tunable ferrite based bandpass filter in the following sections. An electromagnet is used to provide a longitudinal bias magnetic field up to 2000 Oe. The magnitude is then reduced to 800 Oe to 0 Oe with a decreasing step size of 200 Oe.

The s-parameters are recorded and plotted in Fig. 4.2. For ferrite thin films, the transmission will show a very well defined single absorption peak. A wide notch with two main peaks is observed for each non-zero cases due to the multi-mode in its thickness direction.
(a)
The magnetic permeability spectrum under varied magnetic bias is plotted in Fig. 4.3. When there is no magnetic bias, the initial permeability is 11, the value is then dropped to a very low value close to zero at higher frequency, and at last, being close to unity. When magnetic field is applied, the shape of the dispersion is quite different from that under zero magnetic bias. Negative permeability region is observed, which has also been shown in previous work from Tsutaoka, et al. [9]. Two peaks were observed which
was due to the existence of multi-modes associated with the large thickness of the YIG slab. This is also shown in the FMR measurement. (Fig. 4.5)

The initial permeability is reduced under stronger in-plane bias magnetic field. At the same time, the negative part of the real relative permeability is moved forward to higher frequencies when the external magnetic field is increased. We can conclude that when the bias magnetic field was larger than 800 Oe, the permeability was always positive in frequency range of 0 – 4 GHz. In the later text, we will show that the YIG substrate would not support propagation wave in the case when the permeability is negative. Thus, if the designed central frequency is around 3.7 GHz, an in-plane magnetic field of over 800 Oe should be applied to make sure the YIG substrate can always operate in the positive permeability spectra region.
(a)

(b)

63
Fig. 4.3 Measured complex permeability of the ferrite (a) real part, and (b) imaginary part.

4.2.2. Magnetic hysteresis loop

The magnetic hysteresis loop is carried out by the Vibrating Sample Magnetometer (VSM) with a saturation magnetization ($4\pi M_s$) of 2100 Gauss, and a coercivity of 1 Oe. (Fig. 4.4) The saturation magnetic field is 200 – 250 Oe for in plane, and near 1500 Oe for out of plane.

Fig. 4.4 Measured normalized magnetic hysteresis loop of the doped YIG sample.
4.2.3. Ferromagnetic resonant frequency (FMR)

The narrow and FMR spectrum is carried out in our lab by using the system described in [10]. RF source with different frequencies from 2.8 GHz to 3.7 GHz is applied. Resonance peaks under various amplitude of magnetic field is observed, as shown in Fig. 4.5. For a RF source with a certain frequency, e. g. 2.8 GHz, the material show a main resonance peak under a magnetic field of 400 Oe and a second peak at around 500 Oe. This indicates that when magnetic field of 400 Oe to 500 Oe is applied on the material, the material will undergo an absorption band around this frequency range, which will result in an unworkable bandgap in the microwave devices. However, the FMR linewidth is measured less than 50 Oe implies that the material still exhibits a low magnetic loss at off-FMR frequency region.
4.3. Resonator measurement verification

A single pole hairpin resonator at S-band is fabricated on the ferrite substrate in order to verify the permeability measurement. First, a Cu seed layer is deposited onto the ferrite substrate following by a 17 μm copper electroplating process. Next, the resonator is patterned by standard photolithography. Finally, the device is put into 15% Nitride Acid for wet etching.

The transmission coefficient under varied magnetic bias is carried out by our network analyzer. The initial resonant frequency is at 3.72 GHz, though the device is to
be designed as a single resonator, the insertion loss exhibits a low value of 0.8 dB, which show a great potential of implement a low loss ferrite based bandpass filter in this band. When the applied magnetic field is increased to 800 Oe, the resonant frequency decreased to 2.74 GHz with an insertion loss of 1.4 dB. The resonant frequency increased to 3.13 GHz when the magnetic field is increased to 1900 Oe, and the insertion loss decreased to less than 0.2 dB. In this case, the frequency up shift is 390 MHz. The resonant frequency up shifts more slowly at higher external magnetic field because the permeability change of the YIG is getting smaller at higher bias fields. In addition, the frequency shift in this band is not linear because of the non-linearity of permeability tuning by the external bias field. A minimum insertion loss of 0.01 dB is obtained under 1300 Oe.
When the magnetic field is applied between 200 Oe and 600 Oe, the device exhibits an absorption band as predicted. The central resonance frequency is around 3.72 GHz, 3.75 GHz and 4.01 GHz, while the insertion loss increase to 7.33 dB, 8.65 dB, and 13.93 dB, respectively. The magnitude of the transmission and reflection are read from Fig. 4.7, and calculated, as shown in Table 1. The small return loss and high insertion loss indicates that almost 30% of the power has been reflected due to the impedance mismatch while little power is transmitted through. The energy dissipation in the ferrite increases from 51.86% to 67.81%. In other words, instead of band-passing behavior, the
wave transmission is almost prohibited between with an external magnetic field from 200 Oe to 600 Oe, the ferrite substrate will show a negative magnetic permeability from 2.3 to 4 GHz.

The relative permeability reading from Fig. 4.3 at each frequency is -9.4, -10.5, and -12. According to the propagation factor $e^{-|\beta|t}$, the phase constant is an attenuation factor in this case, thus the wave will decay very quickly as the negative permeability value decreased. This matches the measured permeability property of the substrate, where shows negative region above 3.5 GHz when the external magnetic field is applied at 200 Oe 400 Oe, and 600 Oe.
Fig. 4.7 Measured s-parameters (S21 and S11) under 200 Oe, 400 Oe and 600 Oe.

<table>
<thead>
<tr>
<th>Magnetic field (Resonance frequency)</th>
<th>200 Oe (3.72GHz)</th>
<th>400 Oe (3.75GHz)</th>
<th>600 Oe (4.01GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Insertion loss (S21)</td>
<td>7.33dB (18.49%)</td>
<td>8.65dB (13.65%)</td>
<td>13.93dB (4.07%)</td>
</tr>
<tr>
<td>Reflection (S11)</td>
<td>5.28dB (29.65%)</td>
<td>5.82dB (26.18%)</td>
<td>5.51dB (28.12%)</td>
</tr>
<tr>
<td>Power Absorbed</td>
<td>51.86%</td>
<td>60.17%</td>
<td>67.81%</td>
</tr>
</tbody>
</table>

Table 4.1 Loss calculation under 200 Oe, 400 Oe, and 600 Oe.
4.4. Conclusion

In this chapter, a method of measuring the complex permeability using a CPW and a network analyzer is presented. The permeability spectra under varied magnetic field are discussed. The measured permeability spectra show a negative region which prohibits the wave propagation. Therefore, the energy of the RF source dissipated in the material, and the attenuation depends on the magnitude of the negative value. In addition, a resonator is fabricated on the ferrite substrate, the device showed an absorption band gap under a
magnetic field from 200 Oe to 600 Oe, which was in good agreement with the measured permeability spectra.
4.5. Reference


Chapter 5: *Multiferroic BPF on Partially Magnetized Ferrite*

5.1. Motivation

Tunable bandpass filters are widely used in modern RF communication systems with ever increasing demand on insertion loss, tunable range, bandwidth, linearity, size, weight, power efficiency [1]. Ferrite filters, such as yttrium iron garnet (YIG) based filters [2]-[7] and self-biased NiCo-ferrite based filters [8, 9] have been studied and designed for different applications for years because of its low loss tangent, narrow ferromagnetic resonance linewidth, wide operating frequency range and high stability. One of the challenges with these types of ferrite filters or microwave ferrites based devices, however, is that most of these microwave ferrites operate in their fully saturated states with high tuning or bias fields up to several kilo-Oersteds in order to tune the ferromagnetic resonant frequency to the frequency of interests due to their low saturation and low anisotropy fields [4]-[6]. This leads to microwave magnetic devices that are bulky and power consuming.

In this chapter, we start from the design of a magnetically tunable bandpass filter on YIG substrate. A magnetically tunable bandpass filter under a low magnetic bias of 100 Oersteds shows a large tunability of over 6%. The power handling capability of the device under varied magnetic bias is then discussed, and a high power handling capability of 20 dBm is reported. Based on the concepts of partially magnetized ferrite’s large
permeability tuning range under low magnetic biases, a dual magnetically and electrically multiferroic tunable bandpass filter on a nickel ferrite substrate is then presented. Compared to widely used YIG material, nickel ferrites exhibit a larger saturation magnetostriction value which react more rapidly to the mechanical force induced by piezoelectric phase.

5.2. Introduction to multiferroic tunable bandpass filters

Multiferroic composite materials consisting both a magnetic phase and a ferroelectric phase are of great current interests, which offer the possibility of magnetoelectric (ME) coupling, and have led to many novel multiferroic devices [4]-[6]. Compared to conventional tunable microwave magnetic devices that are tuned by magnetic field [3, 7], these dual H- and E-field tunable microwave multiferroic devices are much more energy efficient, less noisy, compact, and light-weight. However, these E-field tunable multiferroic devices typically show very limited tunable frequency range due to their limited effective E-field induced effective magnetic fields, which are typically in the range of 5~50 Oe in epoxy bonded ferrite/ferroelectric heterostructures.

Most recently, Srinivasan, et al. [5] (Fig. 5.1) reported a magnetoelectric microwave bandpass filter with a single crystal yttrium iron garnet–lead zirconate titanate (YIG-PZT) bilayer to implement the communication of two microstrip antennas at
ferromagnetic resonant (FMR) frequency. By applying electric field across the thickness of the piezoelectric sample, the mechanical deformation in piezoelectric phase results in an anisotropy change in the magnetic phase, i.e., the YIG single crystal, which corresponds to a FMR tuning. The metal electrodes on both sides of PZT plate produces an electrical wall on top of the YIG single crystal, the device exhibits a reasonable insertion loss of around 5 dB. Yang, et al. [4] (Fig. 5.2) also reported a better tunability at L-band by using a similar concept yet they did not specify the insertion loss acquired from the electric field tuning approach.
Fig. 5.1 Magnetoelectric microwave bandpass filter. (a) Device schematic and (b) measured S21 parameter.
Figure (a) shows a schematic of a microstrip line structure with dimensions $W_3$ and $S_3$. The microstrip line is labeled with 'GGG 500 um' and 'YIG 100 um'. The ground plane is marked below the microstrip line.

Figure (b) presents the S-parameters ($S_{11}$ and $S_{21}$) as a function of frequency (GHz). The S-parameters are measured for different magnetic field strengths: 50 Oe, 100 Oe, 150 Oe, 200 Oe, and 250 Oe. The frequency range is from 1.5 GHz to 2.5 GHz.
Fig. 5.2 Dual H- and E-field tunable bandpass filter by Yang, et al. [4] (a) Schematic of the bandpass filters, (b) magnetic tunability and (c) electrical tunability under varied magnetic biases.

This type of devices is of importance due to their fast responding time at micron-seconds compared to a level of millisecond’s responding time for traditional magnetically tunable devices. On the other hand, an electrical wall formed by metal electrodes of piezoelectric plates will degrade the insertion loss.
5.3. Magnetically tunable BPF based on partially magnetized ferrite

We present a magnetically tunable bandpass filter on partially magnetized YIG substrate here as a start point. As stated in previous chapters, the permeability of partially magnetized ferrites exhibits a very low value as well as a reasonable magnetic loss when the ratio $\gamma 4\pi M_s/\omega$ is close to unity. Therefore, by carefully selecting the ferrite with a proper saturation magnetization, bandpass filter with an initial central frequency at different band can be implemented. At the same time, the material is also operated in a low loss regime when the frequency is tuned within a limited range. At this point, the need for low magnetic bias will lead the devices that are more power efficient.

5.3.1. Research efforts on BPF using partially magnetized ferrites

Partially magnetized ferrites have been used in tunable devices [10, 11] due to their large permeability tunable range, low or zero bias field, and fast tuning speed at micron-seconds. Analyses on the permeability tensor of partially magnetized ferrites [12]-[16] indicated that these unsaturated ferrites have $0 < \mu < 1$, which can be tuned by changing its magnetization $M$ between $0 \sim M_s$ (the saturation magnetization) with a low bias field. The permeability, at this point, exhibits a very low value as well as a reasonable magnetic loss when the ratio $\gamma 4\pi M_s/\omega$ is close to unity. Therefore, by carefully selecting the ferrite with a proper saturation magnetization, bandpass filter with an initial central frequency at different band can be implemented. As a result, large frequency tunability can be achieved in tunable RF/microwave ferrite devices with
partially magnetized ferrites with a low bias field, e.g. less than 150 Oe, since the operating frequency \( f \propto 1/\sqrt{\mu_r} \). At this point, the material is operated in a low loss regime when the frequency is tuned within a limited range. The need for low magnetic bias will lead the devices that are more power efficient.

Oates, et al. [11] (Fig. 5.3) reported a magnetically tunable superconducting bandpass. The device was realized by printing a microstrip structure on top of a polycrystalline YIG substrate and then tested under a low magnetic bias of less than 300 Oe. The tunability of the devices was 13% and the losses were well below 1 dB in superconducting environments.
Fig. 5.3 Magnetically tunable superconducting bandpass filter (a) device schematic and (b) S21 parameters.

5.3.2. Device construction

The device was constructed by following the same design consideration, but was implemented with a more compact size in order to set the initial resonant frequency at C-band to fulfill the requirement of the operating frequency range with respect to the selected ferrite material. The dimension of the device is 5mm × 7mm with a thickness of 0.3mm. (Fig. 5.4) The ferrite substrate is chosen from the commercial product pure polycrystalline YIG from Trans-Tech Inc (G-113). The saturation magnetization $M_S$ is 1816 Gauss with a coercivity of 1.5 Oe. The YIG ceramic slab has a dielectric constant of
$\varepsilon' = 14.59$, an electric loss tangent $\tan\delta=0.00004$, and a linewidth of less than 24 Oe. The structure of the bandpass filter is a compact, low loss single pole hairpin design consisting of a “U” shaped resonator capacitively coupled to the input and output port. The strip-lines were constructed by standard photolithography with 10-μm thick copper. A uniform magnetic field from 0 to 100 Oe was applied parallel to the feed line. In this case, the ferrite substrate is operated in a unsaturated state.
5.3.3. Simulations and experimental verification

Unlike traditional simulation of magnetic bias excited spin waves inside the magnetic materials, when the magnetic material is used as a substrate in a partially magnetized state, it is hard to simulate the magnetic properties of the material under varied magnetic bias. The HFSS only simulate the magnetic material in saturated states. Therefore, the calculation of magnetic properties of the material is presented as a first step of the simulation.
Following the theory of partially magnetized ferrites discussed in chapter 3, the calculated complex permeability in demagnetized state is \( \mu_r^0 = \mu_{r0} - j\mu_{r0}'' = 0.71 - j0.0053 \) according to eq (3.19). By knowing the permittivity, the electric loss and the complex permeability value, one can predict the performance of the proposed bandpass filter by inserting the values into the material spec boxes. On the other hand, the permeability can also be derived from the equation by knowing the instantaneous magnetization, the magnitude of bias field and the resonant frequency of the bandpass filter.

The \( \mu' \) under varied magnetic biases are calculated from such approach. (Fig. 5.5) The calculated values correspond to a magnetic field sweep from 0 Oe to 100 Oe with a step of 10 Oe in measurement. The simulation is implemented by setting the substrate permeability value from 0.7 to 0.85 with a step size of 0.03. (Fig. 5.6)
Fig. 5.5 Calculated real part of complex permeability under varied magnetic biases in partially magnetized states.

The insertion loss and position of initial central frequency in simulation, at zero magnetic bias, is in agreement with the measurements indicates that the calculations stand in this model. (Fig. 5.7)
Fig. 5.6 Simulated performance of the proposed bandpass filter (a) transmission and (b) reflection.
For non-zero cases, there may exist a difference between simulations and measurements/calculations. For instance, at 100 Oe magnetic bias, the calculation gives a permeability of 0.87 (Fig. 5.5) and 0.85 from the simulation. (Fig. 5.6) An error of two percent is obtained.

(a)
The initial central frequency without a magnetizing field is 6.17 GHz with an insertion loss of 1.1 dB. (Fig. 5.8) The 3-dB bandwidth is about 810 MHz. The bandpass filter shows a large frequency shift of 290 MHz when the magnetizing field is increased to 60 Oe. At this point, the central frequency is decreased to 5.88 GHz with a constant insertion loss of 1.1 dB. The lowest central frequency occurs at 5.79 GHz under a magnetic field of 100 Oe with a reasonable insertion loss of about 1.25 dB. The 3-dB bandwidth is mainly determined by the hairpin structure. Narrower bandwidth can be obtained by creating a multi-pole model. The bandwidth remains almost the same when
the YIG substrate transits from unsaturated to saturated states [2]. Compared to the frequency shift when the bias magnetic field is less than 60 Oe, it shows a smaller shift of 80 MHz. This is due to the low variation of permeability under higher magnetic field beyond 60 Oe [13].

The increase of the insertion loss at lower frequency indicates that with the increase of magnetic field, the operating frequency is approaching the “low-field loss” region of YIG material. The low bias field induced FMR absorption region in this case is at about 1.2 GHz and will not contribute to the insertion loss [17] for this design.
Fig. 5.8 Measured (a) transmission (b) reflection coefficients under different magnetizing fields. The measured central frequency shifted downward from 6.17 GHz to 5.79 GHz.

When strong bias field around 1 kOe is applied, the material is saturated. The permeability spectrum exhibits a negative region which overlaps with the operation band of the bandpass filter [2]. The electromagnetic waves quickly decay in the substrate and will result in a high insertion loss over 10 dB. During the measurement, the return losses were more than 18 dB indicating a good impedance match.
The central frequency simulation and measurement when the bias magnetic field is varied from -100 Oe to 100 Oe is plotted. (Fig. 5.9) The simulated frequency under varied bias magnetic field shows reasonable fit to the measured result. The difference is induced by the error from the estimation of permeability by [8]. Small difference at low permeability value ($\mu<1$) will result in relatively large change in frequency domain. The central frequency shows hysteresis behavior to the magnetic field due to the magnetic hysteresis of the ferrite. This indicates that the magnetization has important effect to the central frequency.

Fig. 5.9 Central frequency of the tunable bandpass filter under magnetic field from -100 Oe to 100 Oe. The measurement (simulation) consists of a forward and a backward loop. The central frequency exhibited a 0.3% shift at zero bias field due to the hysteresis.
Green and Sandy [15] measured the magnetization and magnetic loss of YIG material at varied temperature from 23°C to 200°C. The saturation magnetization will decrease with the increase of temperature. This will compress the tunability of the bandpass filter since the frequency is directly related to the variation of magnetization. In real applications, an additional packaging technique, e.g. heat sinks may be applied to dissipate the heat effectively.

To summarize, the bandpass filter exhibits a large frequency tunability of 380 MHz (6.1%), a low insertion loss at the level of 1 dB and a good impedance match. By varying the length of the hairpin resonator, initial central frequency at different location can be obtained. However, for different initial central frequency, the permeability operating range and the loss of the material varies from on to another depending on the distance between the ratio $\gamma 4\pi M_s/\omega$ and unity. A simulation is given (table 5.1) that depicts the trade-off between these factors. Higher permeability tuning range is obtained and will result in higher frequency tunability. In the meantime, more loss will be introduced from the magnetic loss of the YIG material.
<table>
<thead>
<tr>
<th>Hairpin length (mm)</th>
<th>fc (GHz)</th>
<th>$\gamma 4\pi M_s / \omega$</th>
<th>$\mu_r$ under zero Oe</th>
<th>$\mu_r'$ under 100 Oe</th>
<th>Tunability (MHz)</th>
<th>Insertion loss (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.35</td>
<td>5.9</td>
<td>0.86</td>
<td>0.67-j0.0063</td>
<td>0.83</td>
<td>450 (7.6%)</td>
<td>1.6</td>
</tr>
<tr>
<td>4.1</td>
<td>6.17</td>
<td>0.82</td>
<td>0.71-j0.0053</td>
<td>0.85</td>
<td>380 (6.1%)</td>
<td>1.1 measured 1.44 simulated</td>
</tr>
<tr>
<td>3.8</td>
<td>6.44</td>
<td>0.79</td>
<td>0.74-j0.0045</td>
<td>0.88</td>
<td>290 (4.5%)</td>
<td>1.2</td>
</tr>
</tbody>
</table>

Table 5.1 Comparison of tunability and insertion loss in different designs.

From table 5.1, it is seen that different from traditional ferrite FMR based design, the initial central frequency is mainly set by the geometry of the microstrip structure. Therefore, different microstrip geometry can be employed for specific applications. The frequency tunability is larger when the designed resonant frequency is more closed to $\gamma 4\pi M_s$, and the permeability is operated in a wider range. On the other hand, the material loss will increase to deteriorate the insertion loss. In addition, once the geometry is printed on to the ferrite substrate, the device can only work within a very limited frequency band. However, the exchange of large tunability and low tuning field is still attractive for ferrite devices.
5.3.4. Power handling capability test

The schematic of the equipment set up for high power measurement is illustrated. (Fig. 5.10) The VNA produce an output that was varied from -27 dBm to 0 dBm, which went through a 30 dB power amplifier (HP 83020A) and an isolator before connecting to the device under test (DUT). The isolator and a 30 dB attenuator were used to protect the RF amplifier and the VNA.

The IP_{1dB} was at 18 dBm when the YIG substrate was operating in completely demagnetized state at zero magnetic bias. (Fig. 5.11) The IP_{1dB} degraded at increased bias magnetic field. It is seen the IP_{1dB} point shows a decreasing trend when the magnetic field is increased. When the field was increased to 150 Oe, the IP_{1dB} was at 11.5 dBm, mainly “low-field loss”. The increase of bias field below saturation will cancel the anisotropy field inside the YIG substrate and the loss increases with a decrease of net magnetic field. Domain wall motion mainly contributes the loss in the YIG material at low bias fields [18], which will be diminished under strong bias field of 1000Oe and above. When the ferrite is fully saturated in MSSW based bandpass filters, the power handling shows a high power handling capability of over 30 dBm [7]. In such case, the domain is completely aligned to the bias magnetic field. Over 30dBm power handling capability is also obtained when strong bias magnetic field is applied and the YIG substrate is in saturated state.
Fig. 5.10 Schematic of the high power measurement set up.

Fig. 5.11 Measured power handling capability of the bandpass filter in partially magnetized and saturated states.
5.3.5. **Summary of magnetically tunable BPF**

The concept of tunable bandpass filters based on partially magnetized ferrite are of importance. The experiments have proved that although the magnetic loss of the ferrite in partially magnetized state is claimed to be large and is not suitable to microwave devices, the results still show an encouraging aspect in achieving both a large frequency tunability and a low insertion loss with a very low magnetic bias. The need for low tuning magnetic fields in the range of <100Oe for partially magnetized ferrites provides a unique opportunity for dual H- and E-field tunable RF/microwave multiferroic devices, since the low tuning magnetic field can be provided by the effective magnetic field induced by E-field.

5.4. **Multiferroic tunable bandpass filters**

In this section, a three-pole hairpin tunable bandpass filters designs using partially magnetized nickel ferrite substrate at Ku-band is presented. The unsaturated ferrite substrate operated at a low permeability range from 0.56 to 0.7 under magnetic fields from 0 to 150 Oe. In addition, multiferroic bandpass filters with ferrite/PMN-PT (lead manganese niobate-lead titanate) structure are demonstrated. The bandpass filters have electrically tunability of 270MHz (2.1%) with an electric field of 9kV/cm under a low bias magnetic field of 100 Oe.
5.4.1. Bandpass filter on Nickel ferrite substrate

In this experiment, polycrystalline Ni-ferrite (NiFe$_2$O$_4$) from Trans Tech Inc. commercial products (TT86-6000) is chosen as the substrate. Properties of the material are listed in Table 5.2. The wider linewidth and slightly larger loss tangent of the nickel ferrite results in more loss in microwave devices; however, the larger magnetostriction compared to YIG materials leads to more electrical tunability in multiferroic applications.

In-plane magnetic hysteresis loops are measured with vibrating sample magnetometer (VSM). (Fig. 5.12) The nickel ferrite with a saturation magnetic field of about 300 Oe, is harder than the previously discussed YIG material. But similar as YIG, the magnetization change of the materials also becomes less sensitive to magnetic field when the ratio $M/M_s$ is greater than 0.6, which result in slower permeability change according to eq (3.19), corresponding to a bias magnetic field of 150 Oe.

<table>
<thead>
<tr>
<th>$\varepsilon'$</th>
<th>$\tan\delta_e$</th>
<th>$4\pi M_s$</th>
<th>$\Delta H$</th>
<th>$\lambda_s$</th>
</tr>
</thead>
<tbody>
<tr>
<td>12.17</td>
<td>0.0002</td>
<td>4750 Gauss</td>
<td>149 Oe</td>
<td>~25ppm</td>
</tr>
</tbody>
</table>

Table 5.2 Properties of nickel ferrite.
To meet the prerequisite of a partially magnetized device using nickel ferrite with high saturation magnetization of 4750 Gauss, a tunable bandpass filter using partially magnetized nickel ferrite at Ku-band is demonstrated. The dimension of the device is 6 mm × 3.5 mm with a ferrite thickness of 0.25 mm. (Fig. 13) The structure is a three-pole hairpin design [19] in order to obtain a narrower bandwidth. The device was constructed by using standard photolithography technique with 5 um thick copper. The device was then attached to a fixture with two SMA connectors for measurement. A bias magnetic field from 0 to 150 Oe was applied parallel to the feeding line.

Fig. 5.12 Magnetic hysteresis loop of TT86-6000 nickel ferrite.
Fig. 5.13 Geometry of the bandpass filter. W=2.3mm, S1=1.3mm, S2=0.25mm, S3=0.2mm, S4=0.55mm.

The initial resonant frequency of the bandpass filter is at 14.16 GHz with a 3 dB bandwidth of 690 MHz (4.8%). The insertion loss is 5.5 dB and the return loss is 15 dB due to high loss of the Ni-ferrite material. The central frequency is decreased to 14 GHz with an insertion loss of 5.2 dB under a 50 Oe field. When the bias magnetic field is increased to 100 Oe and 150 Oe, the frequency is decreased to 13.72 GHz and 13.35 GHz, and the insertion loss is 5.5 dB and 6.1 dB, respectively. The 3 dB bandwidth is increased to 950 MHz under 150 Oe and the bandpass behavior deteriorates. The resonant frequency, in this case, is close to the “low-field loss” region and the quality factor of the material is decreased. The return losses of the bandpass filter are greater than 15 dB and the rejection band are more than 25 dB. The return loss exhibits high value below 12 GHz due to the high magnetic loss of the material at this frequency range.
Fig. 5.14 Measured (a) transmission and (b) reflection coefficients under different magnetizing fields.
In order to obtain a large electrically tunability, bandpass filter with multiferroic heterostructures was fabricated at 13.8 GHz. At lower frequency, the material exhibits a higher permeability tuning range. The calculated permeability in demagnetized state is \( \mu_{\text{demag}} = 0.51 - j0.015 \). More energy dissipates in the materials in change of a higher tunability. The frequency tuning range is from 13.8 GHz to 12.89 GHz under a magnetic bias up to 150 Oe. (Fig. 5.15) The insertion loss of this device is around 14 dB (Fig. 5.18) at this point. However, from previous experiments on nickel ferrite device, we had demonstrated that the insertion loss is able to be compressed within 5 to 6 dB at Ku-band. A fabrication improvement will lead to less insertion loss which is more suitable for applications.

![Graph showing frequency distribution versus magnetic bias at a lower frequency.](image)

Fig. 5.15 A frequency distribution versus magnetic bias at a lower frequency.
5.4.2. Ni-ferrite/PMN-PT Multiferroic heterostructure

The multiferroic heterostructures was formed by epoxy bonding a 0.5 mm thick ferroelectric PMN-PT (lead manganese niobate-lead titanate) single crystal, with \( d_{31} \) along x-axis, onto the back side of the device. (Fig. 5.16) The structure produces a strong mechanical coupling and allows voltage tuning of the band pass filter. The glue electrically isolated the device ground plane and the PMN-PT so that the piezoelectric only has mechanical contribution to the filter. The effective magnetic field leads to an in-plane anisotropy change inside the ferrite substrate which refers to a high permeability tuning.

(a)
Compared to the widely used PZT ceramics used in many multiferroic devices, the PMN-PT single crystals with a $d_{31}$ of -1800pC/N and a $d_{32}$ of 900pC/N, not only have a much larger piezoelectric coefficients, but also show an in-plane anisotropic piezoelectric behavior that is critical for achieving large E-field tunability [20].

The nickel ferrite polycrystaline used here has a saturation magnetostriction $\lambda_s \sim$ -33ppm. Ideally, assuming the mechanical coupling between the ferroelectric phase and magnetic phase is lossless, the effective in-plane magnetic field of the Ni-ferrite/PMN-PT bilayer induced by stress can be obtained from inverse magnetoelastic relation (cgs):

$$\Delta H_{eff} = 3\lambda Y d_{eff} E/M_s$$

(5.1)
where \( \lambda \) is the ME coefficient. \( Y \) is the Young’s modulus of bulk nickel spinel (1.68 \( \times \) 10\(^{12}\) dyne/cm\(^2\)), \( E \) is the electric field applied on the PMN-PT/ferrite bilayer, and \( d_{\text{eff}} \) is the effective piezoelectric coefficient of the PMT-PT single crystal is given by

\[
d_{\text{eff}} = \frac{\nu d_{32} + d_{31}}{1-\nu^2} - \frac{\nu d_{31} + d_{32}}{1-\nu^2} = \frac{d_{31} - d_{32}}{1+\nu} \quad (5.2)
\]

where \( \nu \) is the poisson’s ratio of NiFe\(_2\)O\(_4\) (~0.32). When an electric field from -1 kV/cm to 9 kV/cm is applied, (2) gives a calculated \( \Delta H_{\text{eff}} \) change of up to 30 Oe in either \( x \) or \( y \) direction.

### 5.4.3. Measurement verification

An electric field from 1 to 9 kV/cm is applied along the thickness or [011] direction of the PMT-PT slab. A magnetic bias with three different amplitudes, 0 Oe, 75 Oe and 100 Oe, along the \( y \)-axis are investigated, which corresponds to the three observation points in Fig. 5.15. The lowest frequency occurs at 13.05 GHz under an electric field of -1 kV/cm under a bias field of 100 Oe. The central frequency is increased to 13.20 GHz and 13.32 GHz when the applied electric field is 5 kV/cm and 9 kV/cm. An electrically tunability of 270 MHz (2.1%) is obtained under a bias field of 100 Oe and the central frequencies show a good linearity to the applied electric field. By using the same measurement technique, tunability of the central frequencies under the magnetic bias of 75 Oe and 0 Oe are obtained as 160 MHz (1.2%) and 135 MHz (0.9%), as shown in Fig. 5.17. This is due to the lower tuning rate of the device with respect to the magnetic field.
under 75 Oe and 0 Oe compared to that under 100 Oe. In addition, the ME coefficient increase of the nickel ferrite under stronger bias magnetic field contributes to the increase of $\Delta H_{\text{eff}}$ and change of anisotropy which is critical to the magnetization of the ferrite. On the other hand, glue bonding induced clamping effect will reduce the substrate bending and result in lower effective magnetic field compared to the calculation.

The observed frequency tunability, with no bias field, exhibits a opposite tuning trend compared to the case of 75 Oe and 100 Oe. This is due to the PMN-PT induced change of in-plane anisotropy under varied electric fields [20]. When bias field along y-axis is applied, the monotonic change in anisotropy results in the unidirectional shift of the central frequency. The tuning rate is reduced when the applied electric field was further increased due to the saturation of the piezoelectric strain at higher E-fields.
Fig. 5.17 Measured E-field tunable operating frequency range of the bandpass filter under different magnetic bias fields, showing the dual E- and H-field tunability.

Fig. 5.18 S21 curves of the multiferroic device under 100 Oe magnetic bias.
A more complete measurement on the central frequency of the bandpass filter as a function of applied electric fields is depicted. (Fig. 5.19) The measurement consists of a forward loop (-9 kV/cm to 9 kV/cm) and a backward loop (9 kV/cm to -9 kV/cm). A clear hysteresis loop can be observed with a “butterfly” shape. This “butterfly” central frequency resembles the widely observed piezoelectric strain vs. electric field curves for piezoelectric materials [20] and matches the ferroelectric P-E hysteresis loop of PMN-PT single crystal as well. Therefore, as previously discussed, the central frequency hysteresis confirms that the device realized the control of the magnetization of nickel ferrite with piezoelectrics induced strains which show great potential for electrically tunable multiferroic devices.
5.5. Conclusion

In this chapter, starting from the state of art of the multiferroic tunable bandpass filters, a question has been raised: how to implement a tunable bandpass filter that has a larger electrically tunability, low insertion loss and power efficient. An engineering approach that combining both the concepts of partially magnetized ferrite and magnetoelastic coupling is proposed. The idea is to use the multiferroic heterostructure in control of anisotropy of the ferrite to tune the permeability in its rapid tuning region.
Two experiments are presented:

1) A planar compact bandpass filter at C-band on a partially magnetized YIG substrate was demonstrated with a large tunability of 380 MHz (6.1%), a low insertion loss of 1.1 dB to 1.25 dB under low magnetic field of 0 to 100 Oe. The bandpass filter on unsaturated ferrite substrate also showed IP$_{1dB}$ up to 18 dBm and over 30 dBm in saturated state.

2) A Ku-band multiferroic tunable bandpass filter on nickel spinel with an electrically tunability of 270 MHz (2.1%) is presented. The experiment discussed the usage of the spinel’s high magnetostriction in microwave devices. The frequency tunability is closely related to the ferrite permeability operating region. In addition, the measured central frequency under varied electric field showed a “butterfly” behavior due to the magnetic hysteresis of the nickel ferrite substrate in response to the PMN-PT single crystal.
5.6. References


Chapter 6: Voltage Tunable Multiferroic Phase Shifter

Phase shifters are important components in RF/microwave systems. They are widely used in many RF/microwave modules and circuits such as phase discriminators, beam forming networks, power dividers, linearization of power amplifiers, and phase array antennas, etc. Tunable phase shifters allow the application of a single device with multiple functions. Among numerous phase shifters they are always required to be tunable, compact, lightweight, low loss, fast responding time, and power efficient, etc.

In this chapter, we will present a compact voltage tunable multiferroic phase shifter on yttrium iron garnet (YIG)/lead magnesium niobate lead titanate (PMN-PT) heterostructure. The device exhibited a phase shift more than 115° under an electric field of 11 kV/cm, and a reasonable low insertion loss of 3.7 dB compared to the state of art of ferrite based phase shifters. The formation and design consideration of the device will also be discussed. A comparison of E-field tunability between 1-layer PMN-PT structure and 2-layer PMN-PT structure using bending mode is discussed. An over 20% improvement on the phase shift over the proposed frequency range is reported. The device combining with multiferroics and partially magnetized ferrite concepts is able to operate with the absence of external bias magnetic field is more compact and power efficient than traditional multiferroic devices.
6.1. Introduction to phase shifters

Phase shifters are used to change the transmission phase angle (phase of S21) of a network. Ideal phase shifters provide low insertion loss, and equal amplitude (or loss) in all phase states. While the loss of a phase shifter is often overcome using an amplifier stage, the less loss, the less power that is needed to overcome it. Most phase shifters are reciprocal networks, meaning that they work effectively on signals passing in either direction. Phase shifters can be controlled electrically, magnetically or mechanically. The major parameters which define the RF and microwave Phase Shifters are: frequency range, bandwidth \((BW)\), total phase variance \((\Delta \varphi)\), insertion loss \((IL)\), switching speed, power handling \((P)\), accuracy and resolution, input/output matching \((VSWR)\) or return loss \((RL)\), harmonics level.

Most commonly used in electronically scanned antenna arrays the antenna beam can be steered in the desired direction without physically repositioning the antenna by applying a phase shifter to each antenna. When steering the beam, the phase of the elements is adjusted so that individual signals line up at the desired beam-pointing angle \((\theta)\). A total phase shift variation of 360° \((2\pi \text{ rad})\) is often needed to realize the phase tuning of a single radiation element. The simplest way of controlling signal phase is to systematically vary the cable lengths to the elements. Cables delay the signal and so shift the phase. However, this does not allow the antenna to be dynamically steered. For this reason, a tunable phase shifter is highly desired for modern radar systems.
6.1.1. State of art of tunable phase shifters

Different techniques and approaches have been employed in achieving phase shift in RF/microwave components. Currently, most phased array antenna systems rely on ferrite [1-3], MMIC [4], or MEMS [5-7] phase shifters. Moreover, new approaches such as piezoelectric transducers (PET) [8, 9], field effect transistor (FET) switches [10], and ferroelectric phase shifters [11] are drawing more attention to the researchers.

MEMS phase shifters have much faster response speeds (measure in milliseconds), however their major drawback is that they have high losses at microwave and millimeter-wave frequencies. Other disadvantages with MEMS phase shifters is that they have limited power-handling capability (20 dBm) and they may need expensive packaging to protect the movable MEMS bridges against the environment. MMIC phase shifters are blazing fast, they can easily change state in tens of nanoseconds, but power handling is limited to 10 milliwatts. They can also be very expensive, as they are processed on gallium arsenide, not silicon. PIN diodes can also be used to make very low-loss phase shifters, however they are mostly controlled by current. These limitations prevent their applications in mission critical phased arrays, such as high power radars and electronic warfare.
6.1.2. Application of ferrites in tunable phase shifters

Ferrite ceramics are used in tunable phase shifters as well as other RF/microwave components for several decades due to the low loss, high power capability, high resolution and reliability. Roome and Hair [12] reported a ferrite-dielectric phase shifter. Though the device was only focused on the interaction between the wave propagation and the ferrite without any tunability, many latter works on magnetic field tunable phase shifters [13-15, 19, 20] were based on the theory. However, ferrite phase shifters are slow to respond to control signals (often at milliseconds) and cannot be used in applications where rapid beam scanning is required.

A desirable alternative is the latching waveguide ferrite phase shifter [1] that operates at the remanent magnetization for the ferrite element and requires current pulses for switching the magnetization state. Dionne and Oates [2] presented a microstrip phase shifter with a similar concept but with a more compact planar profile. Though a very large phase shifter over 1000°/dB was reported, the device was operated under a superconducting environment. The ferrite in this type of device is operated before saturation, by changing the magnetic flux of a closed loop within the path of ferrite, the ferrite is then magnetized into any states. As previously discussed, a rapid change of permeability will be obtained in this case. On the other hand, this type of devices is not reciprocal and need to be reset in order to be tuned to another new state.
Most recently, Srinivasan [16, 17] reported planar voltage tunable phase shifters with YIG-PZT/PMN-PT heterostructure. A ME element is positioned on top of a designed microstrip structure. (Fig. 6.1a) An in-plane external magnetic bias of 2.7 kOe was applied in order to tune the FMR frequency. The complex permeability then has a steep regime as well as a low loss near FMR as region 1 and 2. (Fig. 6.1b) When an electric field across the PZT is applied, the influence of mechanical deformation on the piezoelectric phase and the magnetic phase will produce a phase shift on the microstrip. The reported electrically phase shift and insertion loss is up to 180° and 3 dB to 4 dB at X-band. (Fig. 6.1c) The phase shift is even much smaller at C-band. On the other hand, the device needs a large magnetic bias of 2.7 kOe which is very power consuming.

(a)
Fig. 6.1 (a) Schematic of the phase shifter with ME element (b) permeability dispersion, and (c) phase shift test result at X-band.

Another voltage tunable multiferroic phase shifter demonstrated by Geiler [18] in also reported a phase shift of 65° with an insertion loss of over 3.2 dB at C-band. Still, this device is also required an external magnetic bias of >200 Oe that makes the whole system lack of spacing efficiency.

Therefore, it will be more competitive if one can realize a phase shifter with large phase shift, low loss, and small profile at the same time.

6.2. Propagation constant in ferrites

The wave propagates on ferrites is quite different from the propagation in widely used dielectrics, such as Rogors materials, etc. The magnetization and the permeability of the ferrite vary from time to time with respect to the external magnetic bias and frequency. Therefore, the propagation constant is set by the properties of the ferrite as well as the microstrip geometry which provided a boundary condition. In 1965, Wheeler obtained an approximation for the impedance and the propagation constant for microstrip line on dielectric substrates. [21] Roome and Hair [22] gave a more detailed calculation of the propagation constant on ferrite in 1968.
For a microstrip line on dielectric when $W/H < 1$, the effective permittivity ($\varepsilon_{eff}$) is given by

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ \left( 1 + 12 \frac{H}{W} \right)^{-1/2} + 0.04 \left( 1 - \left( \frac{W}{H} \right)^2 \right) \right]$$  \hspace{1cm} (6.1)

where $\varepsilon_r$ is the relative permittivity of ferrite, $H$ and $W$ are the thickness of the substrate and the width of the microstrip line, respectively.

The filling factor $q$ can be expressed by

$$q = \frac{\varepsilon_{eff} - 1}{\varepsilon_r - 1}$$  \hspace{2cm} (6.2)

The effective permeability for the transmission line is then a weighted parallel combination of the relative permeabilities:

$$\mu_{eff} = \frac{1}{1 + q \left( \frac{1}{\mu_r} - 1 \right)}$$  \hspace{2cm} (6.3)

where $\mu_r$ is the ferrite relative permeability. When the ferrite is magnetized to saturation and the signal RF magnetic field $h_{rf}$ is parallel to the direction of an in-plane magnetization ($4\pi M$), there is a minimum interaction between the RF magnetic field and the ferrite; the relative permeability is approximately equal to unity, which the fractional change is small. When the angle ($\varphi$) of the magnetization and the RF magnetic field is
90°, there is a maximum interaction. (Fig. 6.2) An approximation of $\mu_r$ in the frequency range $\omega_m/\omega < 1$ under small magnetic bias is given by [22]

$$\mu_r \approx 1 - \left(\frac{\omega_m}{\omega}\right)^2$$

(6.4)

where $\omega_m = \gamma 4\pi M$, $4\pi M$ is the ferrite magnetization, and $\gamma$ is the gyromagnetic ratio being 2.8 GHz/kOe.

Fig. 6.2 Wave propagation, signal RF magnetic field and magnetization configuration.

When the magnetization $4\pi M$ is varied with small external magnetic bias $H$, if the FMR frequency $f_0 \ll f$, the effective permeability in this case is given by [23]
\[
\mu'_{\text{eff}} \approx 1 - q \frac{\gamma^4 \pi M}{f} \left( \frac{\gamma}{f} \right)^{N_y - N_z} \frac{4 \pi M}{f} \]  
(6.5a)

\[
tan \mu = \left| \frac{\mu''_{\text{eff}}}{\mu'_{\text{eff}}} \right| \approx q \left( \frac{\gamma^4 \pi M}{f} \right)^{\frac{\Delta f_0}{f}} \]  
(6.5b)

where \(N_y\) and \(N_z\) are demagnetizing factors along \(y\) and \(z\)-axis, suppose the wave propagates along \(z\)-axis. For a slab with \(y\)-axis perpendicular to the board face, \(N_x = N_z = 0, N_y = 1\). \(\Delta f_0\) is the resonance linewidth from spin-lattice relaxation damping.

By knowing these values, one can obtain the propagation constant in ferrite in terms of frequency \(f\) yields

\[
\beta (f) = 2 \pi f \sqrt{\mu_0 \varepsilon_0 \mu_{\text{eff}} \varepsilon_{\text{eff}}} \]  
(6.6)

where \(\mu_0\) and \(\varepsilon_0\) are magnetic and dielectric constant in free space, respectively. For a microstrip with a fixed line width and substrate thickness, the only parameter can be tuned by the external magnetic bias is the effective permeability \(\mu_{\text{eff}}\). Therefore, the phase shift of a microstrip ferrite phase shifter is approximately estimated by

\[
\Delta \phi = (\beta_{H=0} - \beta_{H>0}) \cdot L \]  
(6.7)

where \(L\) is the effective length of the microstrip along the ferrite magnetization axis.
6.3. Device construction

In this section, a proposal of voltage tunable multiferroic phase shifters will be presented. Different piezoelectric operating mode will also be discussed.

6.3.1. Meander line phase shifter with YIG/PMN-PT bilayer

A phase shifter on polycrystalline YIG substrate (TransTech–G113) was designed and fabricated. (Fig. 6.3) The dimension of the device is 20 mm × 10 mm with a thickness of 0.3mm. The structure of the phase shifter is a compact quarter wavelength meander line design with a length $L_M$ of about 100 mm.

The phase shifter was designed at C-band. The ferrite operating in this frequency range exhibits a low permeability and a relatively low magnetic loss tangent as previously discussed. Small tuning of the magnetic bias will result in high relative variation of the. Therefore, large tunability in phase response is expected.
Fig. 6.3 (a) Geometry of meander line phase shifter and (b) fabricated device on YIG substrate with PMN-PT bonding.
The fabricated phase shifter on YIG substrate is also shown in Fig. 6.3. The strip-lines were constructed by electrode plating with 17 µm copper following by wet etching techniques. The phase shifter was attached to a fixture with an SMA connector during measurement. An in-plane uniform magnetic field was applied perpendicular to the feed line (Fig. 6.3a) in order to obtain a magnetically tunable phase shifter on partially magnetized substrate.

Next, a 0.5mm thick PMN-PT single crystal plate with an approximately equal size was bonded onto the back side of the ferrite substrate using epoxy. (Fig. 6.3b) The glue electrically isolated the ferrite and the PMN-PT plate so that the metal electrodes as well as the applied voltage across the PMN-PT do not affect the electrical performance of the phase shifter. Instead, a deformation of the piezoelectric will produce a mechanical force which will be transferred to the ferrite phase through the ferrite-piezoelectric bilayer.

6.3.2. Phase shifter with stack 2-layer PMN-PT/YIG structure

By knowing the relation between phase shift and the propagation constant from eq (6.6), larger phase shift can be obtained by increasing the length of the microstrip. Therefore, another structure is designed on an identical substrate with a three-stage
cascade double spiral microstrip structure. For each single element, the trace follows the equation (Fig. 6.4a)

\[
x = \pm 0.4mm/\pi \cdot \theta \cdot \sin \theta \tag{6.7a}
\]

\[
y = \pm 0.4mm/\pi \cdot \theta \cdot \cos \theta \tag{6.7b}
\]

where \(\theta\) is from \(2\pi\) to \(6\pi\). The two spirals are then connected with an “S” shaped arc line. The length \(L_s\) of the cascade structure is calculated to be approximately 130 mm. By increasing (decreasing) the spacing between the lines, the center of the operating frequency band is decreased (increased).

The strip-lines were also constructed by electrode plating with 10 µm copper following by wet etching techniques. Then, a 0.5mm thick, (011) cut and (001) poled PMN-PT single crystal PMN-PT single crystal plate was bonded onto the back side of the ferrite using epoxy with \(d_{31}\) along the \(x\)-axis. In order to strengthen the mechanical deformation of the piezoelectric phase, another piece of PMN-PT single crystal with the same size was bonded to the first PMN-PT plate to form a stacked 2-layer structure. (Fig. 6.4b) The second plate was also (011) cut and (001) poled. The \(d_{31}\) was also along the \(x\)-axis. An opposite electric field was applied to the two PMN-PT plates for measurement at this point.
Fig. 6.4 (a) Geometry of the phase shifter and layer demonstration, (b) fabricated device with fixture.
6.4. Simulation

The transmission and reflection coefficients of the two designs are simulated with HFSS V12. Since the ferrite model in HFSS are all in fully saturated case, similar to the simulation techniques in the previous chapter, the permeability is set as a sweeping parameter in order to obtain the coefficients in different states.

It is seen from Fig 6.5 and 6.6, for both designs, the insertion losses are less than 3 dB and the return losses are better than 10 dB. However, in real devices, the insertion loss at this frequency range is expected to be higher than simulation due to the “low field loss” of the ferrite material. Especially for the double spiral structure with a narrower bandwidth, high insertion loss of the material and the cut-off region of the wave propagation will add up to high loss at lower frequencies. However, by increasing the line gaps one can lower the pass band central frequency to obtain higher phase shift and lower loss.
Fig. 6.5 Simulated (a) transmission and (b) reflection coefficients of the meander line ferrite phase shifter.
6.5. Numerical results

The simulation on ferrite in HFSS when the ferrite is in partially magnetized state is quite different from that in fully saturated state which is mostly used in spin wave mode. The ferrite permeability spectrum is a function of frequency and can vary by over 80 percent in a very small frequency interval. In addition, the interaction between the signal RF magnetic field and the magnetization is also need to be taken in to account. Therefore, in this section, the numerical results of the meander line design (Fig. 6.3) is brought up as the simulation result.
The YIG material used in this device exhibited a dielectric constant $\varepsilon_r = 14.59$. Thus, from Eq (6.1), (6.2), one can obtain the effective permittivity $\varepsilon_{eff} = 9.67$, and the filling factor $q = 0.64$. The phase shift of the phase shifter can be obtained from Eq (6.5) – (6.7) as in Fig. 6.7.

![Fig. 6.7 Calculated phase shift of the meander line phase shifter.](image)
Fig. 6.8 A fitted hysteresis loop curve of the ferrite substrate obtained from measured phase shift.

Different from the microstrip phase shifter on commonly used dielectrics, such as Rogor’s materials, the calculated phase shift exhibits a decreased phase shift with respect to the increasing of the frequency, which can also be more precisely calculated from the Galekin’s method in Ref. [15] and [24]. During the calculation, the normalized magnetization value is set to fit with the measured phase shift. In this case, the y-axis and z-axis demagnetizing factor with the value $N_y = 1$ and $N_z = 0$. (assuming the wave propagates along z-axis and y-axis is perpendicular to the ferrite board face). However, in real device, a y-axis effective demagnetizing factor $N_{y_{eff}} < 1$ is introduced since the rf signal intersects the ferrite surface adjacent to the conductor microstrip and forms
magnetic poles. [23] This results in the difference between the fitted hysteresis loop and
the measured magnetic hysteresis loop of the ferrite substrate. (Fig. 6.8)

6.6. Measurement verification

6.6.1. Meander line phase shifter

*Magnetically tunability:*

Transmission and reflection coefficients measurements of the fabricated bandpass
filter were carried out by a network analyzer under various external magnetizing fields
from 0 to 100 Oe. (Fig. 6.9) The insertion loss is 3.6 dB under zero bias magnetic field
from 5.5 GHz to 7.5 GHz. When the bias field is increased to 100 Oe, the insertion loss is
increased to 5.1 dB. The insertion loss is induced mainly by the “low-field” loss. This is
due to the domain wall motion when the ferrite substrate operated in partially magnetized
state. Return losses are all better than 15 dB indicating a good impedance match during
the measurement. It is seen that the insertion loss is increased and the pass band behavior
is deteriorated, at lower frequency, when the applied magnetic field is increased. This is
due to the increasing of cut-off frequency of wave propagation in ferrite substrates under
higher magnetic field.
Fig. 6.9 Transmission and reflection of the phase shifter under varied magnetic bias.

The phase shifter shows a 360° phase shift under a 100 Oe bias magnetic field. (Fig. 6.10) In this case, the magnetization of the ferrite substrate is about 80% saturated according to the magnetic hysteresis loop of YIG material. Increasing the magnetic field beyond 100 Oe will only result in small phase shift. The permeability change will become slower with respect to the magnetic field due to small changes of magnetization. Compared to the calculated results, the phase shift shows a rapid increasing rate for stronger applied field at lower frequencies, this is due to the cut-off threshold of wave propagation is pushed forward to higher frequency under stronger magnetic field, and results in a rapid change of propagation constant.
Fig. 6.10 Measured (colored lines) and calculated (colored scatters) phase shift under varied magnetic bias.

It is seen the device exhibits an over 105° phase shifter under 15 Oe bias field, this gives an opportunity for realizing multiferroic device controlled by electric field since low effective magnetic bias produced by piezoelectric-ferrite heterostructure is suitable for the low bias field requirement in this device.

The measured differential phase shift per insertion loss in degree/decibel under varied external magnetic field is plotted in Fig. 6.11. A 68 degree/dB loss is measured at
6 GHz under low bias magnetic field of 100 Oe. A maximum value of 72 degree/dB is observed at 6.55 GHz, which corresponds to a phase shift of 307 degree. In this case, the phase shifter exhibits a low insertion loss of about 4.25 dB. Table 6.1 shows a specification comparison on phase shift per dB insertion loss of magnetically tunable ferrite phase shifters in listed references. It is seen that the designed phase shifter combined both large phase shift per decibel loss, low loss and low tuning magnetic field. The phase shift per insertion loss was decreased very quickly near the mode of cut off at lower frequency with the increasing of the bias magnetic field.

Fig. 6.11 Phase shift per insertion loss of the phase shifter. An over 72°/dB phase shift is obtained.
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<th>Phase shift/ dB Loss</th>
<th>Bias field (Oe)</th>
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<td>4.25</td>
<td>72 (7.4/cm)</td>
<td>100</td>
</tr>
</tbody>
</table>

Table 6.1 Phase shift per decibel loss comparison of magnetically tunable ferrite phase shifters.

*Electrically tunability:

In order to obtain an electrical tunability, a voltage from -50 V to 200 V was applied over the thickness of the PMN-PT plate, which produces an electric field from -1 kV/cm to 4 kV/cm. The insertion loss of the phase shifter is less than 3.7 dB above 5.5 GHz, which is not affected by the bonded PMN-PT, as predicted. It is seen when the electric field is applied, the pass band at around 5.17 GHz is shifted upward by $\Delta f = 80 \text{ MHz}$, which is mainly caused by the change of magnetic properties of the ferrite. The insertion loss below 5.5 GHz drops very quickly as the same found in magnetic tuning.
case due to the wave propagation cut off condition in ferrite materials. When the voltage across the PMN-PT is beyond 200 V, i.e. 4 kV/cm, the phase shift does not increase indicates that the mechanical deformation of the piezoelectric phase is almost saturated in this case and no mechanical force is transferred to the ferrite phase.
Fig. 6.12 Measured (a) transmission coefficient and (b) reflection coefficient under varied electric field.
Fig. 6.13 Measured differential phase shift of the multiferroic phase shifter with meander line structure.

The phase shifter in this design is much less than the value which predicted in magnetic tuning discussion. There are several reasons for the explanation:

1. The magnetic bias applied to the device is along the width direction, i.e. parallel to most of the microstrips, which produces a maximum phase shift for ferrite substrate based phase shifters. However, the effective anisotropy change induced by ME coupling is not necessarily along longitudinal direction which results in an inefficient use of the microstrip.
2. The stiffness of the ferrite is much more than expected due to the large thickness. The mechanical deformation of the piezoelectric cannot be effectively transferred to the ferrite, which results in less tunability.

### 6.6.2. Double spiral phase shifter

Due to the symmetry of the double spiral structure to the origin, the effective length of microstrip remains almost unchanged to the ferrite magnetization in any direction. The measurement consists of two sub-measurements. First, a PMN-PT plate was bonded to the back side of the ferrite substrate using epoxy. The device was tested under a voltage from -50 V to 550 V, which corresponding to an electric field from -1 kV/cm to 11 kV/cm along the thickness direction. Then, another PMN-PT single crystal with the same size was bonded onto the first PMN-PT using epoxy with the same polarization and $d_{31}$ direction. Electric fields, at this time, with equally magnitude are applied onto both PMN-PT slabs at the same time but are in opposite direction in order to produce more mechanical force. Detailed information can be found in 6.3.2.

Differential phase shift from 5 GHz to 6 GHz was measured for both cases, as shown in Fig. 6.14. When the electric field is increased from -1 kV/cm to 11 kV/cm, the 1-layer PMN-PT phase shifter exhibited a phase shift of 50° at 5.5 GHz and a maximum of 94° in the observed range. This indicates that the magnetization of the ferrite is
changed due to the ME induced anisotropy change of the material, thus results in a large permeability tuning.

Fig. 6.14 Measured differential phase shift of the phase shifter with 1-layer PMN-PT plate.

When a second PMN-PT slab is bonded, the device shows a maximum phase shift of over 115° at 5 GHz and 60° at 5.5 GHz, as in Fig. 6.15. The phase shift was improved by an overall of 20% compared to the 1-layer PMN-PT/ferrite design. The stacked PMN-PT 2-layer structure, in this case, is operated in a bending mode. The opposite applied E-field on the two PMN-PT slabs leads to a tensile stress on one slab, and compressive on the other, thus leads to a shape extension to one plate and a shape compression to the
other one. When the two PMN-PT plates are bonded together, the whole module will produce a bending stress that leads to an enhanced magnetoelectric coupling on the ferrite layer, as illustrated in Fig. 6.15b. Therefore, stronger mechanical force will be transferred to the ferrite substrate compared to the single PMN-PT/ferrite structure. This will result in larger anisotropy change inside ferrite substrate through ME coupling, which will lead to larger permeability tuning when the ferrite is operated at a very low value.
Fig. 6.15 (a) Measured differential phase shift with stacked 2-layer PMN-PT. (b) Stacked 2-layer PMN-PT bending mode demonstration.

It is seen that the phase shift rate is decreased when the applied electric field is beyond 5 kV/cm, and the phase shift increases very slowly when the electric field is greater than 11 kV/cm. This is due to the PMN-PT single crystal induced mechanical force to ferrite substrate is gradually saturated, and the anisotropy change in the ferrite through ME coupling is increased very slowly. It is notable that the phase shifter with stacked PMN-PT/ferrite heterostructure is able to achieve an electrical tunability of over 115° without any external biasing magnetic field.
The simulated and measured S-parameters are plotted in Fig. 6.16. The measured insertion loss is about 3.7 dB to 8.2 dB and the return loss more than 15 dB. Usually, the insertion loss is mainly induced by magnetic loss from the ferrite material; the conductor loss and impedance mismatch. However, in this case, the loss from the ferrite material dominates the insertion loss, which gives a larger insertion loss at lower frequencies. Nonetheless, we have demonstrated a lower insertion loss of less than 5 dB using meander line structure even at lower frequencies above 5 GHz. Therefore, by optimizing the geometry of the double spiral structure, increasing the line gap to bring passband window to lower frequency, it is possible to achieve a large phase shift as well as a reasonable insertion loss overall of 5 dB at the same time.

Fig. 6.16 Measured and simulated s-parameters of the designed phase shifter.
6.7. Conclusion

In this chapter, the wave propagation mechanisms in ferrite have been discussed. The phase shift of a ferrite substrate phase shifter is closely related to the fractional change of the effective relative permeability, which varies with the variation of the angle between the signal RF magnetic field and the direction of ferrite magnetization.

A planar compact phase shifter at C-band on partially magnetized YIG substrate was demonstrated with a large differential phase shift of 350° at 6 GHz, a low insertion loss of 3.6 dB to 5.1 dB under a low magnetic field of 0 to 100 Oe. The phase shift exhibited a maximum of 72 degree/decibel loss at 6.55 GHz. Besides, an initial voltage tunable multiferroic phase shifter is proposed. The measured voltage tunability is less than 40° under a voltage of 200 V, which corresponding to an electric field of 400 kV/cm. The limited tunability is mainly because of the direction of anisotropy tuning through ME coupling is not parallel to the wave propagation, i.e. perpendicular to the signal RF magnetic field. This degraded the coupling between the RF magnetic field and the ferrite thus results in a smaller permeability tuning than expected in magnetic tuning case.

To mitigate the effect of the decoupling, a 3-stage cascaded double spiral shaped structure is employed which exhibits an omnidirectional geometry to the direction of
magnetization. A differential phase shift up to 94° is obtained with the PMN-PT/ferrite heterostructure. Planar electrically tunable multiferroic phase shifter on a stacked 2-layer PMN-PT/ferrite structure was demonstrated with a large differential phase shift over 115° under an electric field of 11 kV/cm. The result is improved by 20% compared to 1-layer PMN-PT design. In addition, the device is able to operate without any magnetic bias leads to devices that are more compact and power efficient. The devices combining with multiferroic and partially magnetized ferrite concept is promising in achieving voltage tunable devices with large tunability.
6.8. References


Chapter 7: Conclusion

In this dissertation, I combined the concept of partially magnetized ferrite and magneto-electric coupling, implemented magnetically tunable resonators, dual electric field and magnetic field tunable bandpass filters, and voltage tunable ferrite phase shifters. The voltage tunable devices are able to operate with the absence of the external magnetic bias which shows great promise for compact, power efficient RF/microwave ferrite tunable devices with large tunability.

A method of measuring the complex permeability using a CPW and a network analyzer is presented. The permeability spectra under varied magnetic field are discussed. The measured permeability spectra show a negative region which prohibits the wave propagation. Therefore, the energy of the RF source dissipated in the material, and the attenuation depends on the magnitude of the negative value. In addition, a resonator is fabricated on the ferrite substrate, the device showed an absorption band gap under a magnetic field from 200 Oe to 600 Oe, which was in good agreement with the measured permeability spectra.

A planar compact bandpass filter at C-band on a partially magnetized YIG substrate was demonstrated with a large tunability of 380 MHz (6.1%), a low insertion loss of 1.1 dB to 1.25 dB under low magnetic field of 0 to 100 Oe. The bandpass filter on
unsaturated ferrite substrate also showed IP_{dB} up to 18 dBm and over 30 dBm in saturated state. A Ku-band multiferroic tunable bandpass filter on nikel spinel with an electrically tunability of 270 MHz (2.1%) is presented. The experiment discussed the usage of the spinel’s high magnetostriction in microwave devices. The frequency tunability is closely related to the ferrite permeability operating region. In addition, the measured central frequency under varied electric field showed a “butterfly” behavior due to the magnetic hysteresis of the nickel ferrite substrate in response to the PMN-PT single crystal.

A planar compact phase shifter at C-band on partially magnetized YIG substrate was demonstrated with a large differential phase shift of 350º at 6 GHz, a low insertion loss of 3.6 dB to 5.1 dB under a low magnetic field of 0 to 100 Oe. The phase shift exhibited a maximum of 72 degree/decibel loss at 6.55 GHz. Electrically tunable multiferroic phase shifters presented a differential phase shift up to 94º with 1-layer PMN-PT/ferrite structure, and over 115º on a stacked 2-layer PMN-PT/ferrite structure operating in bending modes. The result is improved by 20%. The device is able to operate without any magnetic bias leads to devices that are more compact and power efficient.