Novel Microwave Magnetic and Magnetoelastic Composite Materials and Devices

A Thesis Presented
by
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To
The Department of Electrical and Computer Engineering
in partial fulfillment of the requirements for the degree of Doctor of Philosophy in the field of Electrical Engineering
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Abstract

Bulk microwave magnetic materials and devices have been widely used in different RF/microwave devices such as inductors, filters, circulars, isolators, and phase shifters. With the even increasing level of integration of RFIC and MMIC, there is an urgent need for new microwave magnetic thin film materials and new integrated RF/microwave magnetic devices. In this thesis, we have addressed these needs in three different areas: (1) exchange biased ferromagnetic/antiferromagnetic multilayer thin films with enhanced anisotropy fields, (2) magneto-electric heterostructures and devices, and (3) metamaterial multilayers for FMR enhancement, tunability, and plane wave absorption. Metallic soft magnetic thin films have been demonstrated to have high saturation magnetization, large permeability and relatively high self-biased ferromagnetic resonance (FMR) frequencies, showing great promise for applications in integrated RF and microwave magnetic devices. One problem for these metallic magnetic films is however their relatively low anisotropy fields that are typically in the range of 10~30 Oe, which severely limit their application frequency range. In this work, we investigated the exchange coupled ferromagnetic/anti-ferromagnetic/ferromagnetic CoFe/PtMn/CoFe multilayer films. These CoFe/PtMn/CoFe multilayer films showed a significantly enhanced anisotropy field of 160 Oe, which was 5~10 times of that of the FeCo films. In addition, a narrow FMR linewidth of 45 Oe at X-band was achieved in the CoFe/PtMn/CoFe trilayer. The exchange coupling in the ferromagnetic/anti-ferromagnetic/ferromagnetic
trilayers leads to a significantly enhanced anisotropy field that is crucial for the application of metallic magnetic films in integrated magnetic RF/microwave devices. The magnetoelectric coupling of novel YIG/PZT, FeCoB/PZT and FeGaB/PZT multiferroic heterostructures were investigated at DC and at microwave frequencies.

An electrostatically tunable band-reject filter device was demonstrated, which had a peak attenuation of greater than 50 dB, 40dB rejection band of 10 MHz, and pass band insertion loss of < 5 dB at ~4.6 GHz. Metamaterial wave absorbers were designed and simulated in HFSS. It involves utilizing a multilayer structure that isolates the electric coupling from the magnetic coupling to absorb radiation from an incident electromagnetic plane wave. The absorber achieved a maximum absorptivity of 46%, at ~4GHz.
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Chapter 1. Introduction

1.1. Magnetic films and materials for microwave integrated devices and applications

Microwave technology encompasses a broad range of applications, largely because of the uniqueness of the electromagnetic (EM) wave properties in this frequency spectrum. The microwave spectrum may be categorized into three sub-spectrums: a high rf frequency range which spans from 300 MHz to 1 GHz, a millimeter wave range 30GHz (10 mm) to 300GHz (1 mm) [1.1], and a traditional microwave range representing the frequencies in-between. This translates to EM wavelengths comparable in size to objects targeted by radar, allowing for the use of high gain antennas (gain being proportional to the electrical size of the antenna). Microwaves are also capable of inducing atomic particle resonance of conductive substances such as water, which is utilized in such applications as cooking, medical treatment and detection, and remote sensing [1.2, 1.3]. However, an extensive and growing application of microwave technology is in the area of communication. This is due to the ability of microwaves to carry large capacities of data due to broader signal bandwidth compared to lower rf frequencies (2kHz to <300MHz) [1.1].

In a basic communication system information is transmitted from a source point to a receipt point some distance away, without any loss of information. However, the laws of physics and limitations of manufacturing inhibit the reproducibility of outgoing information at the point of receipt. The collective
properties of microwaves therefore make microwave technology well suited for communications. It is used to achieve wider bandwidth, higher operating speeds, and lower interference due to lower signal crowding. In addition to better radar resolution due to smaller wavelengths, one can produce smaller transmitter/receiver (transceiver) systems, due to smaller component size. One example is handheld cellular phones.

Line of sight travel is a property of microwaves in which the EM wave is not bent by the ionosphere (as would be the case at lower rf) [1.1]. As a result, communication by microwaves must be designed with this limitation in mind. Therefore, radio towers and satellites form an important framework in global and local communications. This opens up the possibility of higher signal losses due to obstructions or atmospheric absorption, and thus the need for active devices such as amplifiers, phase shifters, and noise filters to maintain or restore the quality of the transmitted and received signals.

Microwave magnetic materials serve both as an enabling technology and means of enhancing microwave device performance. An enabling technology is a technology that opens the door to functionality not previously available. There are two major categories of microwave magnetic materials, and these are ferrites and metallic magnetic alloys. Ferrites are ferrimagnetic materials that are particularly attractive at microwave frequencies because they do not obey the reciprocity principle, and can control relatively high powers. This allows microwave systems to exhibit different impedance and phase characteristics depending on the direction of energy flow, with a high degree of thermal and drift stability.[1.4] Table 1.1-1 provides an
overview of the chemical composition of ferrites. Ferrites are polycrystalline (and sometimes single crystal) ceramic materials that are typically formed using a high temperature sintering process [1.4-1.7]. They exhibit dielectric behavior with high resistivity and low dielectric losses as shown in Table 1.1-2, and are widely used in microwave and millimeter wave applications such as inductors, circulators, isolators, phase shifters, switches, tunable resonators and filters [1.8]. Metallic magnetic alloys are ferromagnetic materials, with very high magnetization compared to ferrites (Table 1.1-2). Another attractive feature is that they typically have low coercivities, allowing them to be utilized without need of an external magnetic bias field to saturate the sample, meaning that they are self-biased. They are metallic in nature, so have lower resistivities (than ferrites) and suffer from conductive losses such as skin effect and eddy currents. This severely limits their applications to rf < 1GHz. However, these materials can be deposited using a low temperature rf sputtering process, making them compatible to semiconductor fabrication. A common application of metallic magnetic alloys is rf planar inductors [1.9, 1.10].

Table 1.1-1. Overview of Ferrite Materials [1.11, 1.12]

<table>
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<tr>
<th>Synthesis</th>
<th>Example</th>
<th>Name</th>
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<tr>
<td>( Fe ) substituted with a divalent metal (eg. manganese, magnesium, nickel, copper, cobalt, zinc, and cadmium)</td>
<td>( (Fe^{++}O, Fe^{+++}<em>{2}O</em>{3}) )</td>
<td>nickel ferrite (NFO)</td>
</tr>
</tbody>
</table>

\[ \downarrow \]

\( (Ni^{++}O, Fe^{+++}_{2}O_{3}) \)
<table>
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<tr>
<th>Fe substituted with two different divalent metals at the same time (e.g., nickel-zinc, nickel-cobalt, nickel-aluminum ferrites)</th>
<th>((\text{Fe}^{++}O, \text{Fe}^{+++}_2O_3))</th>
<th>nickel-zinc ferrite (NZFO)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(\downarrow)</td>
<td>((\alpha\text{NiO}, \beta\text{ZnO}, \text{Fe}_2O_3))</td>
<td></td>
</tr>
<tr>
<td>(\alpha + \beta = 1)</td>
<td></td>
<td></td>
</tr>
<tr>
<td>(\text{Fe}_2O_3) substituted with another trivalent metal, (e.g. aluminum)</td>
<td>((\alpha\text{NiO}, \beta\text{ZnO}, \text{Fe}_{(2-x)}\text{Al}_xO_3))</td>
<td>nickel-zinc-aluminum ferrite</td>
</tr>
<tr>
<td>(\downarrow)</td>
<td>((\alpha\text{NiO}, \beta\text{ZnO}, \text{Fe}_{(2-x)}\text{Al}_xO_3))</td>
<td></td>
</tr>
</tbody>
</table>
| Garnets: substitution of rare earth metal (e.g., samarium, gadolinium, dysprosium, holmium, erbium, ytterbium, etc.). | \((3\text{M}_2\text{O}_3,5\text{Fe}_2\text{O}_3)\) | yttrium ferrite (YFO)
| | \(\downarrow\) | or |
| | \((3\text{Y}_2\text{O}_3,5\text{Fe}_2\text{O}_3)\) | yttrium garnet (YIG) |
| Hexaferrites: 3 Oxides combined to compositionally form an M, W, Y, Z, or S type hexaferrite. | \((\text{BaO}), (\text{MeO}), (\text{Fe}_2\text{O}_3)\) | Barium Hexaferrite (BaM) |
| | \(\downarrow\) | |
| | \(M = (\text{Ba}, \text{Fe}_{12}\text{O}_{19})\) | |
| | \(W = (\text{BaMe}_{2}, \text{Fe}_{16}\text{O}_{27})\) | |
| | \(Y = (\text{Ba},\text{Me}_{2}, \text{Fe}_{12}\text{O}_{22})\) | |
| | \(Z = (\text{Ba},\text{Me}_{2}, \text{Fe}_{25}O_{41})\) | |
| | \(S = (\text{Me}_2, \text{Fe}_2\text{O}_8)\) | |
Table 1.1-2. Material Properties of Microwave Magnetic Materials. [1.4, 1.9, 1.10, 1.11, 1.13]

<table>
<thead>
<tr>
<th>Material Properties</th>
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<th>Metallic Magnetic Alloys</th>
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<tr>
<td>Resistivity [Ω⋅cm]</td>
<td>10e6 to 10e9</td>
<td>10e-6 to 50e-6</td>
</tr>
<tr>
<td>Relative permittivity</td>
<td>10 to 14</td>
<td>~1</td>
</tr>
<tr>
<td>$\tan \delta \times 10^4$</td>
<td>2 to 200</td>
<td></td>
</tr>
<tr>
<td>Relative permeability</td>
<td>&lt;100</td>
<td>10 to 1000</td>
</tr>
<tr>
<td>$4\pi M_s [G]$</td>
<td>&lt;7000</td>
<td>10,000 to 24,500</td>
</tr>
<tr>
<td>$H_k [Oe]$</td>
<td>90 to several 100s</td>
<td>10 to 100</td>
</tr>
<tr>
<td>$\Delta H$</td>
<td>1 to 1000 (wide range)</td>
<td>1 to 1000 (wide range)</td>
</tr>
<tr>
<td></td>
<td>(single crystal YIG ~0.2)</td>
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</table>

1.2. Monolithic microwave integrated circuits (MMICs)

Microwave integrated circuits fabricated on a single substrate are known as monolithic microwave integrated circuits (MMICs). The acronym “MMIC” however has become synonymous with microwave and millimeter wave integrated circuits in general, in which active and passive components as well as transmission line distribution networks, are fabricated on a single semi-insulating semiconductor substrate [1.14]. The trend from microwave integrated circuits performing a specific rf function (such as switching, or amplifying) to multifunctional integrated circuits is motivated by the desire for low cost portable appliances. Prior to MMICs, microwave integrated circuits were assembled by combining active and passive microwave ICs.
and discrete components to achieve a desired system function [1.15]. Semiconductor batch processing was used primarily to fabricate high speed transistors and diodes which were wire bonded to a microwave substrate along with other components. This approach is still used today for higher level system assembly. As microwave ICs became more commercialized by the growth of industries such as wireless communications (Table 1.2-1), cost became an important driver [1.15]. This drove the need to efficiently produce higher volumes of product to offset fabrication costs. High speed semiconductor microfabrication satisfies the need for high volume manufacturing because this approach can simultaneously produce multiple copies of a device on a single substrate; the more copies on a substrate and the larger the substrate the higher the product throughput (die count \(=\) die footprint/wafer diameter).

<table>
<thead>
<tr>
<th>Application</th>
<th>System</th>
<th>Frequency Spectrum</th>
</tr>
</thead>
<tbody>
<tr>
<td>cellular communications</td>
<td>Transceiver circuits</td>
<td>0.8-5GHz</td>
</tr>
<tr>
<td>point-to-(multi)point radio</td>
<td>power amplifiers, receivers</td>
<td>25-60GHz</td>
</tr>
<tr>
<td>automotive radar</td>
<td>Transceivers</td>
<td>77GHz</td>
</tr>
</tbody>
</table>

Material and manufacturing costs continue to drive system integration and size. As such, process compatibility plays an important role when introducing new technology to enhance performance of MMIC devices and applications. As an example, III-V semiconductor materials such as GaAs, GaN, and InP are utilized to meet microwave performance objectives, not achievable with the more mature silicon
based materials. Yet, IC fabricators leverage the manufacturing equipment and
techniques of silicon technology to produce higher volumes of parts to offset the
higher cost of these III-V materials. These less mature semiconductor materials
continue to increase in substrate diameter to increase the number of devices produced
per wafer (substrate) batch; further reducing unit costs of the end product. High
temperature syntheses used to grow ferrites restrict their use in semiconductor
fabrication. In addition, the need for an external magnetic bias to saturate the ferrites
also restricts the compact-ability of ferrite components which impacts cost. These
issues have motivated research into developing self-biased ferrites [1.16], as well as
low temperature synthesis techniques such as spin-spray deposition [1.17]. Spin-
spray deposition is a very promising process in that it is more compatible to
semiconductor manufacturing techniques; however there are still challenges in
developing self-biased materials, in addition to materials with the same level of
quality as ferrites grown using traditional methods [1.18]. Metallic magnetic alloys
are lossy due to conductive losses, and have limited operating range (<1GHz) due to
low FMR. The introduction of antiferromagnetic materials in metallic magnetic
multilayers has been demonstrated to enhance the operating frequency range [1.9].
This report will examine the use of trilayer ferromagnetic/antiferromagnetic/ferromagnetic layered composites to increase the
operating frequency range of metallic multilayer thin film composites.

The continual drive for lower cost portable appliances with less power
consumption has changed the dynamics of what a MMIC is. By packing more off-
chip functions into a single system-on-a-chip (SOP), to cut down on assembly costs, MMIC functionality and circuit density has increased pushing the limits on heat dissipation, and feature size. Multilayer (or 3-D) MMIC architectures have been developed to overcome these issues [1.15]. In a multilayer MMIC, the die footprint can be reduced or maintained by extending SOP functionality vertically using interconnecting circuitry between multiple IC layers (Figure 1.2-1). An additional benefit of this approach is that the layers need not be made of the same substrate material, allowing for optimal design and modularity of sub-functions within the MMIC stack. For example, ferrite devices may be fabricated on an MgO substrate layer, while transistor circuitry could be fabricated on a semiconductor layer. The layered architecture however, creates new challenges in both electrical interconnectivity and isolation between layers. Physical interconnects using microvias are limited by its feature size and it may not be possible to form a hole through a substrate layer [1.15], so an alternative method of interconnectivity is to use electromagnetically coupled interconnects. Two common methods are capacitive coupling, and parallel plate wave guides with slot vias [1.15]. Conducting planes are used for interlayer electromagnetic isolation. The drawback of using conducting planes is that they may actually contribute to noise by exciting parallel plate modes resulting in power loss or cross talk between neighboring circuits [1.15(N.K. Das, p.83)]. This report will explore the use of metamaterials for narrowband noise suppression and filtering, and propose an apparatus for characterizing plane wave absorption and filtering in metamaterial multilayer structures.
The added functionality of single layer and multilayer MMICs also creates new opportunities for on-chip tunability. Schaumann and Karsilayan [1.14(p.199)] describes the basic criteria for on-chip tunability in MMICs as follows: (1) the surface area of the control circuitry must be small compared to that of the device to be tuned (eg. filter), i.e. the circuitry must be relatively simple so that semiconductor real estate is conserved. (2) The tuning circuitry must not generate excessive thermal noise or signal noise (eg. cross-talk or modulation) that can reduce the dynamic range of the filter. (3) Power consumption of the tuning circuitry should be small so as to minimize the overall power consumption, (particularly important for battery-operated
systems). This report will examine the use and characterization of magneto-electric layered composites for electrostatic and magnetostatic tunability.

1.3. **Ferromagnetic/antiferromagnetic/ferromagnetic (FM/AFM/FM) trilayers**

Polycrystalline metal magnetic thin films are being actively explored for applications in rf/microwave devices, such as magnetic band stop filters and magnetic integrated inductors, primarily due to their high saturation magnetization and low-temperature synthesis. Magnetic thin films that are suitable for microwave applications typically need to have excellent magnetic softness with a uniaxial anisotropy field and a low coercivity. The magnetic softness desired for rf/microwave applications is often associated with a relatively low anisotropy field, less than 10–20 Oe for most polycrystalline metal magnetic films. The low anisotropy fields of these metal magnetic thin films correspond to a low ferromagnetic resonance (FMR) frequency $f_{\text{FMR}}$, as described by the well-known Kittel equation:

$$f_{\text{FMR}} = \sqrt[2]{\frac{\gamma}{4\pi M_s + H_k}} \quad \text{(cgs units)}$$

with $\gamma$ being the gyromagnetic constant of 2.8 MHz/Oe, $4\pi M_s$ the saturation magnetization, and $H_k$ being the effective anisotropy field of the magnetic thin films. The typical FMR frequency of the soft magnetic thin films is less than 1–2 GHz, which severely limits their applications.

Research studies on exchange coupled layered composites include CoO/Co bilayers [1.21, 1.22], IrMn/Co bilayers [1.23], FeF$_2$/Fe bilayers [1.24], MnF$_2$/Fe
bilayers [1.24], MnPd/Fe bilayers [1.25], FeMn/Fe bilayers [1.26], NiMn/NiFe bilayers [1.27], Fe$_{50}$Mn$_{50}$/Co bilayers [1.28], PtMn/NiFe bilayers [1.29], NiFe/NiO bilayers [1.30], as well as Fe/Cr/Fe trilayers [1.31], FeNi/Cu/FeMn trilayers [1.32], NiFe/FeMn/NiFe trilayers [1.33], FeNi/Cr/FeMn trilayers [1.32, 1.34], and Co/FeMn/CuNi trilayers [1.35]. Exchange-coupled IrMn/CoFe multilayers in particular have been shown to significantly boost the FMR frequency to over 5 GHz [1.36, 1.37]. The common observation is that ferromagnetic/antiferromagnetic (FM/AFM) bilayer thin-film layered structures show enhanced effective anisotropy fields due to exchange coupling, which introduces an additive field ($H_{ex}$). $H_{ex}$ being the exchange bias field which can be expressed as $H_{ex}=J_{ex}/(M_s t_F)$ with $J_{ex}$ being the interfacial exchange energy between the FM and AFM layers. The enhanced effective anisotropy field and the improved FMR frequencies of bilayer FM/AFM thin films along with a single domain state with an ~100% squareness ratio ($M_r/M_s$) is desired for many microwave device applications [1.38, 1.39]. However, exchange coupling in the FM/AFM bilayer thin film structure can also lead to an unwanted increase in coercivity of the FM layer.

Strong exchange coupling typically leads to enhanced ferromagnetic linewidth in exchange coupled multilayer films. A monotonic drop of ferromagnetic resonance (FMR) linewidth as well as a drop in the effective anisotropy field was typically observed in exchange coupled FM/AFM bilayers with the increase in thickness of the FM layer (shown in Figure 1.3-1), which is undesired for real applications. And this would generally hold true provided that the FM layer thickness is much less than the
skin depth at the operating frequency of interest, after which eddy current losses begin to kick in (which would typically be in the range of 1000Å). Furthermore, it has been observed that the measured exchange bias field at dc can differ from that measured at microwave frequencies in exchange coupled AFM/FM bilayers.

Figure 1.3-1. Previous work showing monotonic drop in exchange bias field and FMR linewidth with ferromagnetic layer thickness in FM/AFM bilayers.

Compared to AFM/FM/AFM trilayers and FM/AFM bilayers, trilayers of FM/AFM/FM have their advantages for many microwave applications. First, trilayers of FM/AFM/FM have a higher effective magnetization $M_{s,\text{eff}}$, which can be expressed
as $M_{s,\text{eff}} = \frac{\Sigma t_{\text{FM}} M_s}{\Sigma t_{\text{FM}} + \Sigma t_{\text{NM}}}$, with $M_s$ and $t_{\text{FM}}$ being the saturation magnetization and thickness of the magnetic layers and $t_{\text{NM}}$ being the nonmagnetic layer thickness, such as AFM layer, etc., and therefore, a higher flux conduction capability. Second, FM/AFM/FM trilayer leads to lower coercivity compared to the bilayers of AFM/FM, which is possibly due to magnetic charge compensation at the magnetic film edges [1.40].

FMR linewidth ($\Delta H_{\text{FMR}}$) of magnetic materials is a parameter of paramount importance for rf/microwave applications such as microwave band stop filters [1.41, 1.42] and rf inductors [1.43-1.45]. Large FMR linewidth leads to a reduced quality factor and increased insertion loss, which are among the major problems associated with the microwave band stop filters. Significant progress has been made on understanding the FMR behavior of exchange-coupled FM/AFM bilayers and its physical contribution to FMR linewidth. However, a relatively less amount of work has been done on exchange-coupled FM/AFM/FM trilayers, and their microwave performances are not well understood.

The magnetic and microwave properties of FM/AFM/FM trilayers as well as multilayers of [FM/AFM/ FM/seed/ dielectric]$_n$ consisting of trilayers with ferromagnetic layers of Co$_{90}$Fe$_{10}$ or Co$_{84}$Fe$_{16}$, and an antiferromagnetic layer of Pt$_{30}$Mn$_{50}$ on different seed layers were examined. Results show that the FMR performances of the FM/AFM/FM trilayers are significantly different from those of the FM/AFM bilayers. The magnetic properties of the CoFe/PtMn/CoFe trilayer thin films at dc and at microwave frequencies were examined and it was shown that a
significantly enhanced anisotropy field and a low FMR linewidth can be achieved simultaneously, which corresponds to zero or a very low exchange bias field at the FMR frequency. One such example is the Ru-seeded CoFe\PtMn\CoFe sandwich structure, which showed excellent magnetic softness with a low hard axis coercivity of 2–4 Oe and a significant enhancement of in-plane anisotropy: 57–123 Oe.

1.4. Microwave magnetoelectric (ME) materials

Layered magnetoelectric (ME) composites have drawn a lot of attention due to the strong achievable ME coupling, which can be used in different rf/microwave devices such as electrostatically tunable bandpass filters [1.46], bandstop filters [1.47], resonators [1.48], phase shifters [1.49], etc. These layered ME composites typically contain at least one magnetic layer (or phase) and one piezoelectric layer (or phase). The elastic coupling between these two phases leads to the tuning of dielectric polarization under an applied magnetic field or the control of an induced magnetization under an external electric field, thus enabling an effective energy conversion between electric and magnetic fields as depicted in Figure 1.4-1. There are two ways to excite the ME composites, either by a magnetic field or by an electric field. The first way to excite an ME composite is by an ac magnetic field at low frequencies (typically 5–10 MHz). This can be done either by a sinusoidal magnetic field or by an impulse magnetic field. The second way to excite an ME composite is by an electric voltage, which is essentially an inverse magnetoelastic effect typically
monitored at microwave frequencies. An applied electric field-induced deformation in the piezoelectric phase leads to deformation of the magnetic phase and therefore a change of the effective magnetic field which is reflected in a shift of the ferromagnetic resonance frequencies.

These ME devices are commonly based upon bulk ME composite materials, in which strong ME coupling has been achieved. Three commonly researched categories of such composites are two-phase ferrite/piezoelectric ceramic structures, two-phase metallic magnetic alloy/piezoelectric structures, and three-phase Terfenol-D/piezoelectric ceramics/polymer structures.[1.50] In order to achieve strong ME coupling at microwave frequencies, the magnetic materials in ME composites need to have a large saturation magnetostriction constant ($\lambda_s$) and high permeability, i.e., a low saturation magnetic field ($H_s$), high saturation magnetization ($4\pi M_S$), and narrow FMR linewidth ($\Delta H$). High quality single crystal yttrium iron garnet (YIG) material has been utilized in microwave ME composite materials and devices, but are not well
suited for MMICs due to high synthesis temperatures mentioned earlier. In addition, YIG has a very low $\lambda_s$ of 0.2 ppm, which is not ideal for achieving strong ME coupling.

An alternative microwave magnetic material for ME composites is the class of metallic magnetic films, of which the eddy current loss at microwave frequencies can be negligible when the thickness of the film is less than its skin depth. These metallic magnetic films can have large $\lambda_s$, relatively low $\Delta H$, high $4\pi M_s$ of up to 24.5 kG, high squareness of ~100%, high self-biased FMR frequencies in the gigahertz range, and low processing temperature.

Metallic magnetic films with excellent magnetic softness and low $\Delta H$, however, typically have very low $\lambda_s$. For example, several of the most well known soft magnetic films such as Permalloy (Ni$_{81}$Fe$_{19}$ wt %), Sendust (FeAlSi), CoCrTa, CoZrNb, FeXN films, etc., all have nearly zero $\lambda_s$.

Soft magnetic thin films based on Fe$_{70}$Co$_{30}$, such as the (Fe$_{70}$Co$_{30}$)$_{100-x}$N$_x$ [1.51] and (Fe$_{70}$Co$_{30}$)$_{100-x}$B$_x$[1.52] exhibit excellent soft magnetic properties with low $\Delta H$, low $H_k$, together with a decent $\lambda_s$ of 40–45 ppm. Most recently, we developed FeGaB films with a narrow $\Delta H$ of 15 Oe at x-band (10 GHz), a low $H_k$ of ~20 Oe, and a large $\lambda_s$ of 70 ppm. Both FeCoB/PZT and FeGaB/PZT composite materials were studied.

Strong ME coupling was observed at microwave frequencies showing large FMR frequency shifts of 50–110 MHz at ~2.3 GHz. When the PZT is subjected to a transverse voltage, deformation of the PZT will lead to strain in the metallic magnetic films (FeCoB or FeGaB) which are tightly bonded to the PZT. This PZT induced
strain in the magnetic film will lead to a change in its in-plane anisotropy field, which is reflected in the shift in the FMR frequency of the magnetic film at microwave frequencies.

### 1.5. Metamaterial multilayer composites

A material may be described macroscopically in terms of its electric response \( \varepsilon \) and magnetic response \( \mu \) to an electromagnetic (EM) wave passing through it. These constitutive parameters of permittivity \( \varepsilon \) and permeability \( \mu \) are related to the phase velocity \( v_p \) of the EM wave as it propagates through the material, where,

\[
v_p = \frac{1}{\sqrt{\mu \varepsilon}} = \frac{c}{\sqrt{\mu, \varepsilon}}.
\]

In 1968, Russian scientist Viktor Veselago [1.53] proposed the concept of a material in which both the permittivity and permeability are negative as shown in Figure 1.5-1. The material would exhibit left-handed (LH) polarization or negative index behavior when an EM wave travels through it [1.53, 1.54]. This gave birth to the idea of using artificial structures (called metamaterials) to produce this LH handed behavior, since no naturally occurring LH materials have been found to exist.

Itoh, et. al. [1.54] defines an electromagnetic metamaterial as a structure whose structural average cell size \( p \) is much smaller than the guided wavelength \( \lambda_g \). Homogeneity is achieved by ensuring that the cell size (and lattice spacing) is at least smaller than a quarter of the guided wavelength \( p < \frac{\lambda_g}{4} \) to minimize EM wave scattering and diffraction effects. The unit cells are typically composed of conducting
elements (copper, gold, etc.) that are arranged periodically to form the metamaterial structure (Figure 1.5-2). The structure behaves as a real material within the limit that
\[ \frac{\lambda_g}{p} > 4 \] or larger. LH metamaterials were first demonstrated by Pendry in 1999, with the development of an unbalanced split ring resonator (SRR) unit cell combined with a wire structure [1.55]. The SRR unit cells were periodically spaced at an average distance \( (p) \). In 2000, Smith and Padilla, et. al. [1.56] improved on Pendry’s design with a balanced (or paired) split ring resonator (SRR) unit cell. A balanced SRR differs from unbalanced SRR in that only the magnetic component of the EM wave couples to the balanced SRR structure, whereas both the electric \( (\vec{e}) \) and magnetic \( (\vec{h}) \) component of the EM wave couple to the unbalanced SRR. SRR based metamaterials may be thought of as having artificial magnetic dipole moments since they can exhibit magnetic behavior within a narrow bandwidth. Over the last several years much of the research in metamaterial structures has gone into development and study of LH materials [1.53, 1.55, 1.56, 1.57], for various applications such as EM wave cloaking, perfect lenses, etc. [1.55, 1.57, 1.58]. Recently, efforts have been made in developing metamaterial EM wave absorbers particularly for the terahertz frequency regime where naturally occurring wave absorbing materials don’t exist [1.59]. This research is specifically geared towards its use in bolometers [1.59-1.61]. There has also been similar work done in the microwave regime [1.60]. Wave absorption may be expressed in terms of the transmitted and reflected power from an EM wave incident on the material, using scattering parameters as
\[ A = 1 - \left| S_{11} \right|^2 - \left| S_{21} \right|^2, \]
where \( A \), \( \left| S_{11} \right|^2 \), and \( \left| S_{21} \right|^2 \) are the absorptivity, reflectivity, and transmissivity.
respectively. Complete absorption is achieved when $|S_{11}|^2 = |S_{21}|^2 = 0$, such that $A = 1$.

For zero reflectivity the wave impedance of the material must be matched to the media of the incident wave.

<table>
<thead>
<tr>
<th>$\varepsilon&lt;0$, $\mu&gt;0$</th>
<th>$\varepsilon&lt;0$, $\mu&lt;0$</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Right-Handed (RH)</strong></td>
<td><strong>Left-Handed (LH)</strong></td>
</tr>
<tr>
<td><strong>Metals at optical frequencies</strong></td>
<td><strong>Material</strong></td>
</tr>
<tr>
<td><strong>Evanescent wave.</strong></td>
<td><strong>Backward wave propagation.</strong></td>
</tr>
<tr>
<td>(Veselago’s Materials)</td>
<td>(Ferrimagnetic Materials)</td>
</tr>
<tr>
<td></td>
<td><strong>Evanescent wave.</strong></td>
</tr>
</tbody>
</table>

Figure 1.5-1. Constitutive Properties of Materials.
For an EM plane wave incident on an absorber as shown in Figure 1.5-3, the impedance matching condition requires that 

\[ Z = \sqrt{\frac{\mu}{\epsilon}} = Z_0 \sqrt{\frac{\mu_z}{\epsilon_z}} = Z_0. \]

This requires that \( \epsilon \) and \( \mu \) of the absorbing material have equal value. If the permittivity and permeability of the material can be independently tuned, the impedance can be matched to any incident media, at any frequency across a narrow bandwidth and polarization. This motivated research into developing a suitable electric resonator structure that couples only to the electric component (\( \vec{E} \)) of the EM wave, to complement the magnetic resonating properties of the balanced SRR [1.59, 1.60, 1.62, 1.63, 1.64]. Figure 1.5-4 provides a summary of the basic magnetic and electric resonators developed, and their circuit equivalents. Padilla, et. al.[1.60] demonstrated...
one of the first metamaterial plane wave absorbers in the microwave spectrum. The electric resonator consists of a pair of gapped capacitors with a stripe inductor in the center (Figure 1.5-4c, 1.5-4d and Figure 1.5-5). The electric field couples to the capacitors when \( \vec{e} \) is aligned perpendicular to the gaps. Surface currents induced along the capacitive legs generate antiparallel magnetic fields which cancel out so that \( \mu = \mu_0 \) in the structure. The stripe inductor of the electric resonator and the metal strip on the bottom layer (called a cut-wire) is used to couple \( h \). As \( h \) passes between the two layers, antiparallel currents are induced as shown in Figure 1.5-6. With this design, Dr. Padilla’s group was able to achieve narrowband absorptivity of 88% at \( \sim 11.5 \) GHz. Smith, et. al [1.62] reported on an alternative structure involving a pair of stripe inductors in parallel with a gapped capacitor (Figure 1.5-4d). This works in a similar manner to the Padilla structure in that the magnetic fields in the structure are cancelled out by surface currents induced by the capacitor. This structure absorbs less electric energy than the Padilla structure because of the smaller effective capacitance. Attempts to narrow the gap or widen the capacitor has the effect of reducing the coupling strength [1.62], so Dr. Smith proposed a structure composed of rectangular spirals to increase the effective inductance so that the resonance frequency of the structure could be reduced (Figure 1.5-7). The beauty these electric and magnetic LC resonator unit cells, is that they are self-oscillating [1.62] in that the coupling effect of each cell is independent of the adjacent cells in the lattice, making the metamaterial less prone to losing its bulk properties along its edges. This paper proposes that the coupling strength of these structures could be increased
by increasing both the effective inductance and capacitance elements. By increasing the absorption density of the unit cell, it may be possible to reduce the scale of the unit cell or maintain the scale of the unit cell at a lower resonance frequency. The inductive stripe elements would be replaced with spiral inductors and the gapped capacitor elements would be replaced with metal-insulator-metal (MIM) capacitors. This paper will explore the feasibility of this approach.

Figure 1.5-3. Plane wave incident on a wave absorber.
Figure 1.5-5. Padilla EM planewave absorber.

Figure 1.5-6. Magnetic coupling in Padilla wave absorber.
Figure 1.5-7. Smith ELC: add turns to increase inductance.

1.6. References

1.14. Y. Sun, *Design of high frequency integrated analogue filters* (IEEE Press, United Kingdom, 2002).
2.1. Thin film deposition

All metallic magnetic films used in this research were deposited using dc or rf magnetron sputtering. RF magnetron sputtering is a physical vapor deposition (PVD) technique in which a substrate is placed between two electrodes in a low-pressure (~10^{-7} Torr) vacuum chamber. The electrodes are driven by an rf power source, which generates a plasma and ionizes the rare gas (such as argon) between the electrodes (anode and cathode). A dc bias voltage is used to provide the positively charged gas ions with enough kinetic energy to drive towards the surface of the cathode, which contains a target metal (such as Fe), and knock off atoms from the target. The displaced atoms condense on the substrate surface to form a film. A strong magnetic field is applied to contain the plasma near the surface of the target to increase the deposition rate. An rf sputtering system may have more than one target, which allows for simultaneous deposition of different metals to form alloys, along with the formation of various layered patterns.

The FM/AFM/FM trilayer films were deposited by Alex Zeltser at Hitachi Global Storage Technologies in San Jose California. Details of sample composition are provided in the sample preparation section of this paper. All other metallic magnetic films were deposited in-house using a PVD rf sputtering system manufactured by AJA International as shown in Figure 2.1-1. The basic apparatus
consists of a vacuum chamber, six rf sputter sources, a rotating substrate holder and a pumping system. A computer interface is used to program the target source, shutter sequence, and sputtering power. The vacuum system contains a mechanical pump, a turbo pump, throttle valve, and chamber.

![PVD RF Sputtering System](image)

Figure 2.1-1. PVD RF Sputtering System.

### 2.2. Sample preparation

FM/AFM/FM trilayers of $\text{Co}_{90}\text{Fe}_{10}/\text{Pt}_{50}\text{Mn}_{50}/\text{Co}_{90}\text{Fe}_{10}$ with a 30 Å Ru seed layer and 30 Å Ru cap layer (referred to as $\text{Co}_{90}\text{Fe}_{10}[\text{Ru}]$ in this context) and $\text{Co}_{84}\text{Fe}_{16}/\text{Pt}_{50}\text{Mn}_{50}/\text{Co}_{84}\text{Fe}_{16}$ with a 30 Å Ru seed layer and 30 Å Ru cap layer (referred to as $\text{Co}_{84}\text{Fe}_{16}[\text{Ru}]$) were deposited. The basic structure is depicted in Figure 2.2-1. To compare the seed layer effects, multilayers of $\text{Co}_{84}\text{Fe}_{16}/\text{Pt}_{50}\text{Mn}_{50}/\text{Co}_{84}\text{Fe}_{16}$
with a 30 Å NiFeCr seed and cap layer (referred to as Co$_{84}$Fe$_{16}$[NiFeCr]) were also deposited. All the ferromagnetic CoFe layers ($t_{FM}$) were varied from 10 to 500 Å, while the AFM layer Pt$_{50}$Mn$_{50}$ remained fixed at 120 Å.

Figure 2.2-1. Depiction of single period trilayer structure with Ru seed layer.
Two sets of eight period trilayers (Co$_{90}$Fe$_{10}$[Ru] and Co$_{84}$Fe$_{16}$[Ru]) alternated with Al$_2$O$_3$ were deposited with a fixed CoFe layer thickness of 200 Å. The Al$_2$O$_3$ dielectric layers were 100 Å thick and were used to suppress eddy current loss. Magnetic-field annealing was carried out for these films to induce the unidirectional anisotropy field by exchange coupling before characterizing these films.

Microstructures of selected FM/AFM/FM trilayer films were characterized using x-ray diffraction (XRD) with a Cu $K\alpha$ source. The Cu source strikes the plane of the test sample at an angle $\theta_{\text{XRD}}$ from $\theta$ to $2\theta$, to produce a two-dimensional pattern of diffracted peaks [2.1]. The peaks correspond to spacing between crystal planes within the sample which is used to determine the crystallography of the sample.

YIG/PZT bilayer ME composites were formed by epoxy bonding 100-µm thick yittrium iron garnet (111) film (YIG), which was grown on a 200-µm-thick substrate of gadolinium gallium garnet (GGG), to a silver-coated 500-µm-thick lead zirconate titanate (PZT) layer. The YIG layer has a dimension of 6x7 mm x 100µm.

FeCoB films were co-deposited from two targets of Fe$_{70}$Co$_{30}$ and B on 0.1mm thick glass substrates. The Fe$_{70}$Co$_{30}$ target deposition power was fixed at 23 W while the B deposition power was varied for achieving different FeCoB alloy films. Figure 2.2-2 shows the coercivity and FMR linewidth as a function of B deposition power. At boron deposition of over 90W, the FeCoB films change from the BCC to amorphous phase, which is accompanied with a low coercivity of <1 Oe, and a narrow FMR linewidth of <20 Oe. FeCoB films in this study were deposited at 90W on the boron target to form a composition of Fe$_{75}$Co$_{15}$B$_{10}$. Details of the FeGaB film
deposition is described in reference [2.2]. The FeGaB films in this study were sputter deposited on 0.24mm Si substrates and have a composition of Fe$_{72}$Ga$_{10}$B$_{18}$.

The metallic magnetic ME composites were formed in a similar manner as the YIG/PZT bilayers. 50 nm thick FeCoB films on 0.1 mm thick glass substrates were epoxy bonded to a 0.5 mm thick Pb(Zr,Ti)O$_3$ ceramic beam with a dimension of 35 x 4 mm$^2$ (PIC151, PI Ceramic Co.). 50 nm thick Fe$_{72}$Ga$_{10}$B$_{18}$ films on 0.24 mm Si substrates were also epoxy bonded to 0.5 mm thick Pb(Zr,Ti)O$_3$ beams of the same dimension.
2.3. Low frequency electrical and magnetic characterization

2.3.1. Vibrating sample magnetometer (VSM)

A vibrating sample magnetometer (VSM) is an instrument that is used to measure the magnetostatic properties of magnetic films. A VSM apparatus generally consists of an electromagnet, piezoelectric mechanical oscillator, pick-up coil transducer, a lock-in amplifier, guassmeter, and controller. The test sample is inserted between the poles of an electromagnet and subjected to dc magnetic field. A piezoelectric oscillator physically vibrates the sample at a fixed sinusoidal frequency. The motion of the sample relative to the dc field modulates the magnetic flux detected by a pick up coil. The pick up coil converts the modulated signal into a voltage that is proportional to the magnetic moment of the sample. A lock-in amplifier is used to measure the voltage relative to the piezoelectric oscillation. A guassmeter measures the applied dc magnetic field. The measurement is synchronized using an instrument controller that is usually driven by a computer interface. The VSM can be configured to measure magnetization, and hysteresis of the sample.

A Lakeshore VSM system shown in Figure 2.3-1 was used to measure in-plane hysteresis along the hard axis (HA) and easy axis (EA) directions to the pinned direction of the multilayer samples. Magnetic fields such as coercive fields, exchange coupling fields, etc., were all measured with a VSM with an error of <1 Oe. The Effective anisotropy fields of these magnetic films were measured by extrapolating
the hard axis minor hysteresis loops (50% saturated), a standard method for extracting anisotropy fields for magnetic materials.

2.3.2. Electric polarization

The dc electric polarization of the ME composites were measured using a precision P-E hysteresis looper (Radiant Technologies, Inc.) shown in Figure 2.3-2. The polarization is determined from measuring electric charge as a function of a swept voltage. The hysteretic effects of the test sample can be measured by looping the sweep voltage.
2.3.3. **Low frequency ME characterization** (ME coupling coefficient)

The low-frequency ME voltage coupling coefficient of an ME composite can be measured under a sinusoidal magnetic field excitation at different bias fields using the apparatus depicted in Figure 2.3-3. The ME composite is inserted in a toroidal coil inductor connected to a voltage oscillator that is fixed at a frequency $f_0$ (typically between 1Hz to 10MHz). The sample is oriented so that the coil generates an in-plane ac magnetic flux that stimulates the magnetic phase of the composite along the easy axis. The composite is also subjected to an easy axis dc magnetic field that is swept along the range $\pm H_{\text{sat}}$, where $H_{\text{sat}}$ is a field that saturates the magnetic phase of the composite. As the dc field is swept, the magnetic phase transitions back and forth.
between a saturated state and a demagnetized state. The ac magnetic field excitation results in a change in the magnetostriction of the magnetic phase which couples to the piezoelectric phase and induces a strain. The strain in the piezoelectric phase generates a transverse charge which is measured as a voltage by an oscilloscope which is connected to the electrodes of the ME composite. The ac excitation magnetic field amplitude and measured electric voltage responses are compared to determine the ME coupling coefficient which may be expressed as 

$$\alpha_{ME} = \frac{\delta V}{\delta H} = \frac{v_p / t_p}{h_m},$$

where $v_p$ and $t_p$ are the induced ac voltage and thickness of the piezoelectric phase, and $h_m$ is the induced ac excitation magnetic field on the magnetic phase. The ME response may then plotted as a function of the dc field.

Figure 2.3-3. Low frequency ME measurement apparatus.
2.4. Narrowband microwave electrical and magnetic characterization

Electron paramagnetic resonance (EPR) is a phenomenon in which the spin of an unpaired electron in a material substance resonates with enough energy to absorb or emit electromagnetic radiation as defined by the relation \( h \nu = \gamma \mu_B B_o \) [2.3]. Where \( h \) is planck’s constant, \( \nu \) is the resonant frequency, \( \gamma \) is the Lande g-factor, \( \mu_B \) is the Bohr magnetron, and \( B_o \) is the magnetic field strength. The presence of unpaired electrons is a property common to magnetic materials. When an external field is applied to a magnetic film the spin of an unpaired electron may either orient parallel or anti-parallel to the external field. A parallel orientation is associated with a lower energy state relative to anti-parallel orientation. The separation between energy states is defined by the relation \( \Delta E = \gamma \mu_B B_o \). This energy separation (or splitting effect) is known as the Zeeman Effect [2.4]. The number of electrons with a parallel orientation (lower energy state) versus the number of electrons with anti-parallel orientation (higher energy state) may be described statistically by the Maxwell-Boltzmann distribution [2.5], as

\[
\frac{n_{\text{higher}}}{n_{\text{lower}}} = e^{\frac{\Delta E}{kT}} = e^{\frac{h \nu}{kT}},
\]

where \( \frac{n_{\text{higher}}}{n_{\text{lower}}} \) is the ratio of paramagnetic centers in the higher energy state to the number of paramagnetic centers in the lower energy state, \( k \) is Boltzmann’s constant, and \( T \) is the temperature in Kelvins. This relationship indicates that the electron spin is generally parallel to the applied magnetic field. Therefore, EPR is usually detected from monitoring EM energy absorption of the tested sample.
2.4.1. **Narrow band FMR detection and monitoring**

A microwave EPR spectrometer test apparatus generally consists of a waveguide resonating cavity, an rf power source, an electromagnet, a guassmeter, a recorder (or plotter), and a controller to drive the instruments. This apparatus can be used to measure relative absorption intensity, $g$-factor, and the FMR linewidth of the sample under test. The waveguide cavity is designed to resonate at a narrow frequency band. The rf source supplies EM radiation to the cavity and monitors the reflected power. The waveguide cavity is oriented between the poles of an electromagnet so that it can be subjected to an external magnetostatic field. A guassmeter is used to measure this applied dc field. A recorder registers the change in reflected power relative to the applied dc field and plots the spectrum of the signal. The test sample is suspended within the cavity by a holder which can be adjusted to orient the sample relative to the dc field. The measurement is synchronized by the controller which can be driven by an instrument panel or computer graphical interface.

The sweep range of the dc field can be determined from the frequency of the rf source using the relation

$$H_o = \frac{h \nu}{g \mu_B},$$

where $g$ is initially assumed to be 2.00 [2.6], and the sweep range is initially set to $-\frac{1}{2}H_o$ to $+\frac{1}{2}H_o$, then adjusted accordingly to trace out the full FMR spectrum.

The sensitivity of the apparatus is related to the minimum quantity of electron spins detected ($N_{\text{min}}$). This quantity is described by the relation,

$$N_{\text{min}} = \frac{k_1 V}{Q_o k_1 v^2 \sqrt{P}},$$

where $k_1$ is a constant, $V$ is the volume of the sample, $P$ is the RF power supplied to
the cavity, and $Q_0$ and $k_f$ are the unloaded quality factor and filling coefficient of the waveguide cavity, respectively [2.3]. This relationship may be approximated by $N_{\text{min}} \sim v^{-\alpha}$, where $0.5 \leq \alpha \leq 4.5$, depending on the spectrometer characteristics, resonance conditions, and sample size (typically $\alpha \sim 1.5$) [2.3]. The FMR linewidth ($\Delta H$) may be determined from the $dP/dH$ spectrum, by measuring the difference in applied dc field between the minimum and maximum of the trace.

2.4.2. Field sweep x-band ferromagnetic resonance (FMR)/electron paramagnetic resonance (EPR) facility (spectrometer)

FMR measurements of the FM/AFM/FM multilayer films were performed at $\sim 9.5$ GHz (x-band) using a Varian E-Line Century Series EPR spectrometer (Figure 2.4-1) with both dc magnetic field and microwave excitation field in the plane of the multilayer samples. In the case of FM/AFM/FM multilayer films, only samples with $t_F$ at or above 50 Å were measured due to weak FMR absorption signal in the FMR spectra. The samples were inserted into a waveguide cavity x-band resonator fitted with a stub tuner to maximize EM coupling. The sample holder consists of a plastic rod with a flat tip. The sample is mounted to the tip with thermo-grease, and suspended in the cavity by inserting it through and opening at the top. The EPR spectrometer was upgraded with an analog-to-digital controller which is driven by a CPU graphical interface. The dc magnetic field ($H_{DC}$) sweep settings and activation was set by computer. During the $H_{DC}$ sweep, the CPU display and EPR plotter simultaneously traces out the $dP/dH$ FMR response. The test samples were manually
rotated in-plane with respect to $H_{DC}$, in an orientation referenced from the exchange axis ($\theta = 0^\circ$) and back ($\theta = 360^\circ$) to observe the effect of exchange anisotropy on FMR response.

2.5. Broadband microwave electrical and magnetic characterization

Compared to the microwave narrowband cavity excitation for the ME materials, broadband air-gap microstrip and co-planar waveguide excitation for ME materials allows for a broadband frequency sweep and a strong magnetic material microwave coupling due to the strong microwave driven magnetic fields.

2.5.1. Microstrip measurement technique

Microstrip Technique
The microstrip FMR test apparatus consists of a coaxial microstrip test fixture (Figure 2.5-1), HP8510 vector network analyzer (with 3.5mm coaxial cables) (Figure 2.5-2), GMW 5403 electromagnet (Figure 2.5-3), and Lakeshore 410 guassmeter. The network analyzer is configured to sweep the frequency from 500MHz to 5GHz, using a minimum of 801 sweep points. A full 2-port calibration is performed at the ends of the coaxial cables of the network analyzer using a 3.5mm coaxial calibration kit. This has the effect of moving the reference plane of measurement from the ports of the VNA instrument panel to the external ports of the cable. The sweep averaging is then set to 32 points. The microstrip test fixture is securely mounted between the poles of the electromagnet using an aluminum mounting block, so that the transmission line is parallel to the poles. The ports of the test fixture are centered on the magnetic poles to ensure maximum field intensity and symmetry. The coaxial cables are then attached to the test fixture and the $S_{21}$ transmission response is measured by setting the VNA to Log-Mag format. The transmission spectrum should have a continuous roll-off and the flatness should not exceed -3dB, in order to minimize any signal distortions in the measurement. The guassmeter probe should then be centered on the fixture to monitor the applied dc magnetic bias. Stability of the test setup is checked by sweeping the external dc magnetic field from 0Oe to 2kOe, and back to 0Oe. There should be little or no change in the frequency spectrum in response to the magnetic field. The sample is coated with a thin coat of 1827 photoresist and inserted between the signal line and ground plane of the test fixture. The magnetic field is set to the reference value $H_{\text{REF}} \geq 25H_{\text{KDC}}$ (where the DC
Anisotropy field $H_{KDC}$ is determined from VSM measurement. The complex full 2-port S-parameters are measured and stored in a text file. The magnetic field is reduced to the highest bias point to be measured ($H_{bias1}$), and the full 2-port S-parameters are measured and stored in a text file. The magnetic field is then reduced to next highest bias point ($H_{bias2}$), and the full 2-port S-parameters are measured and stored in a text file. The measurement sequence is repeated until the S-parameters for all magnetic bias points are measured. The FMR spectrum for each bias point is plotted relative to the saturation spectrum as,

$$\text{TransmissionLoss}[dB] = 20 \log_{10} \left| \frac{S_{21(H_{sat})}}{S_{21(H_{eff})}} \right|.$$  

The anisotropy field ($H_K$) is determined from the known sample $4\pi M_S$ by plotting the FMR frequency versus the DC magnetic bias. The FMR linewidth ($\Delta f_{\text{FMR}}$) is determined from the 3dB bandwidth (3dB-BW) of the FMR spectrum.

Figure 2.5-1. Depiction of rf fields in air-gap microstrip. [2.7]
Figure 2.5-2. (Right) 8510B vector network analyzer. (Left) Agilent E8364A vector network analyzer.

Figure 2.5-3. GMW 5403 electromagnet.

**ME measurement**

ME composites were measured by attaching voltage test leads to the sample. The microstrip ME test apparatus consists of an 8510 vector network analyzer,
coaxial microstrip test fixture, GMW 5403 electromagnet, Lakeshore 410 gaussmeter, MPJA 9305-PS DC power supply, TREK model 609B voltage amplifier, and Fluke 79 Series II multimeter. The biasing apparatus is shown in Figure 2.5.4. The network analyzer is configured to sweep the frequency from 500MHz to 5GHz, using a minimum of 801 sweep points. A full 2-port calibration is performed at the ends of the coaxial cables of the network analyzer using a 3.5mm coaxial calibration kit. This has the effect of moving the reference plane of measurement from the ports of the VNA instrument panel to the external ports of the cable. The sweep averaging is then set to 32 points. The microstrip test fixture is securely mounted between the poles of the electromagnet using an aluminum mounting block, so that the transmission line is parallel to the poles. The ports of the test fixture are centered on the magnetic poles to ensure maximum field intensity and symmetry. The coaxial cables are then attached to the test fixture and the $S_{21}$ transmission response is measured by setting the VNA to Log-Mag format. The transmission spectrum should have a continuous roll-off and the flatness should not exceed -3dB, in order to minimize any signal distortions in the measurement. The gaussmeter probe should then be centered on the fixture to monitor the applied DC magnetic bias. Stability of the test setup is checked by sweeping the external DC magnetic field from 0Oe to 2kOe, and back to 0Oe. There should be little or no change in the frequency spectrum in response to the magnetic field. The sample is coated with a thin coat of 1827 photoresist and inserted between the signal line and ground plane of the test fixture. With the voltage supply set to zero volts, the voltage leads are connected to the test leads of the sample.
Stability of the test setup is checked again by sweeping the external DC magnetic field from 0Oe to 2kOe, and back to 0Oe. Once again, there should be little or no change in the frequency spectrum in response to the magnetic field. The voltage is set to the maximum reverse value \((-V_{\text{max}})\), with the voltage monitored with the voltage meter. The magnetic field is set to the reference value \(H_{\text{REF}} \geq 25H_{\text{KDC}}\) (where the DC anisotropy field \(H_{\text{KDC}}\) is determined from VSM measurement). The complex full 2-port S-parameters are measured and stored in a text file. The magnetic field is reduced to the highest bias point to be measured \((H_{\text{bias1}})\), and the full 2-port S-parameters are measured and stored in a text file. The magnetic field is then reduced to next highest bias point \((H_{\text{bias2}})\), and the full 2-port S-parameters are measured and stored in a text file. The measurement sequence is repeated until the S-parameters for all magnetic bias points are measured. In order to capture the hysteretic effect of the electric polarization of the sample, the magnetic bias measurements are repeated for \(-0.5V_{\text{max}}\), \(-0.25V_{\text{max}}\), \(0\ V_{\text{max}}\), \(0.25V_{\text{max}}\), \(0.5V_{\text{max}}\), \(V_{\text{max}}\), \(0.5V_{\text{max}}\), \(0.25V_{\text{max}}\), \(0\ V_{\text{max}}\), \(-0.25V_{\text{max}}\), \(-0.5V_{\text{max}}\), and \(-V_{\text{max}}\). The FMR spectrum for each bias point is plotted relative to the saturation spectrum as, \(\text{TransmissionLoss}[\text{dB}] = 20\log_{10} \left| \frac{S_{21|H_{\text{bias}}}}{S_{21|H_{\text{ref}}}} \right|\). The anisotropy field \((H_{K})\) is determined from the known sample \(4\pi M_s\) by plotting the FMR frequency versus the DC magnetic bias. The FMR linewidth \((\Delta f_{\text{FMR}})\) is determined from the 3dB bandwidth (3dB-BW) of the FMR spectrum. The FMR shift due to electrostatic bias is determined from the change in FMR frequency for each fixed magnetic bias. The ME coupling field \((\Delta H_E)\) is derived from the expression \(\Delta f_{\text{FMR}} = \gamma \sqrt{\mu} \Delta H_E\), where \(\gamma\)
is the gyromagnetic constant 2.8MHz/Oe, and $\mu_i$ is the relative initial permeability derived from $\mu_i = \frac{4\pi M_s}{H_K} + 1$.

Figure 2.5-4. ME Biasing scheme: a) Voltage Power Supply drives the amplifier using b) junction box between voltage supply and c) High voltage amplifier input. The output of the amplifier connects to d) junction box between amplifier output and ME test fixture.

2.5.2. **CPW measurement technique**

**g-CPW Technique**

The grounded coplanar waveguide (g-CPW) FMR test apparatus consists of an Agilent E8364A vector network analyzer, coaxial g-CPW test fixture, and GMW
5403 electromagnet, and Lakeshore 410 gaussmeter. The network analyzer is configured to sweep the frequency from 500MHz to 5GHz, using a minimum of 801 sweep points. A full 2-port calibration is performed at the ends of the coaxial cables of the network analyzer using a 3.5mm coaxial calibration kit. This has the effect of moving the reference plane of measurement from the ports of the VNA instrument panel to the external ports of the cable. The sweep averaging is then set to 32 points. The g-CPW test fixture is securely mounted between the poles of the electromagnet using an aluminum mounting block, so that the transmission line is parallel to the poles. The ports of the test fixture are centered on the magnetic poles to ensure maximum field intensity and symmetry. The coaxial cables are then attached to the test fixture and the $S_{21}$ transmission response is measured by setting the VNA to Log-Mag format. The transmission spectrum should have a continuous roll-off and the flatness should not exceed -3dB, in order to minimize any signal distortions in the measurement. The guassmeter probe should then be centered on the fixture to monitor the applied DC magnetic bias. Stability of the test setup is checked by sweeping the external DC magnetic field from 0Oe to 2kOe, and back to 0Oe. There should be little or no change in the frequency spectrum in response to the magnetic field. The sample is coated with a thin coat of 1827 photoresist and place on the transmission line of the test fixture like the one shown in Figure 2.5-5. The magnetic field is set to the reference value $H_{REF} \geq 25H_{KDC}$ (where the DC anisotropy field $H_{KDC}$ is determined from VSM measurement). The complex full 2-port S-parameters are measured and stored in a text file. The magnetic field is reduced to the highest
bias point to be measured \( (H_{\text{bias1}}) \), and the full 2-port S-parameters are measured and stored in a text file. The magnetic field is then reduced to next highest bias point \( (H_{\text{bias2}}) \), and the full 2-port S-parameters are measured and stored in a text file. The measurement sequence is repeated until the S-parameters for all magnetic bias points are measured. The FMR spectrum for each bias point is plotted relative to the saturation spectrum as, \( \text{TransmissionLoss[dB]} = 20 \log_{10} \left| \frac{S_{21H_{\text{sat}}}}{S_{21H_{\text{ref}}}} \right| \). The anisotropy field \( (H_K) \) is determined from the known sample \( 4\pi M_S \) by plotting the FMR frequency versus the dc magnetic bias. The FMR linewidth \( (\Delta f_{\text{FMR}}) \) is determined from the 3dB bandwidth \( (3dB-BW) \) of the FMR spectrum. The permeability spectrum for each magnetic bias point is determined from the relation, \( \mu_r = Z_o \frac{1 + S_{11H_{\text{sat}}} - S_{21H_{\text{sat}}}}{1 - S_{11H_{\text{sat}}} - S_{21H_{\text{sat}}}} \) \( j(k_f l \mu_o \omega_o) \) \( (2.8) \), where \( l \), \( t \), \( \mu_o \), \( \omega_o \), and \( Z_o \) are the sample length, sample thickness, permeability constant, angular frequency, and characteristic impedance of the fixture respectively. A scaling factor \( k_s \) \( (0.1 \leq k_f \leq 1 \) , typically \( k_f=0.15 \) ) is determined from curve fitting and extrapolation of the relative permeability back to \( \mu_r = \frac{4\pi M_S}{H_K} + 1 \) at the bias condition \( H_{\text{bias}}=0Oe \), and frequency~0Hz. A permeability spectrum is shown in Figure 2.5-6.
Figure 2.5-5. Grounded coplanar waveguide (gCPW) test fixture used for thin film broad band characterization (500MHz to 5GHz).

Figure 2.5-6. Broad band permeability spectrum.
ME measurement

The microwave permeability ME response was measured using the grounded CPW test fixture shown in Figure 2.5-7. The ME sample is attached to a sample holder which applies a voltage bias to the clamped end of the sample using a voltage supply and amplifier. The grounded coplanar waveguide (g-CPW) ME test apparatus consists of an Agilent E8364A vector network analyzer, coaxial g-CPW ME test fixture (Figure 2.5-8), GMW 5403 electromagnet, Lakeshore 410 gaussmeter, MJPA 9305-PS DC power supply, TREK model 609B voltage amplifier, and Fluke 79 Series II multimeter. The network analyzer is configured to sweep the frequency from 500MHz to 5GHz, using a minimum of 801 sweep points. A full 2-port calibration is performed at the ends of the coaxial cables of the network analyzer using a 3.5mm coaxial calibration kit. This has the effect of moving the reference plane of measurement from the ports of the VNA instrument panel to the external ports of the cable. The sweep averaging is then set to 32 points. The g-CPW test fixture is securely mounted between the poles of the electromagnet using an aluminum mounting block, so that the transmission line is parallel to the poles. The ports of the test fixture are centered on the magnetic poles to ensure maximum field intensity and symmetry. The coaxial cables are then attached to the test fixture and the $S_{21}$ transmission response is measured by setting the VNA to Log-Mag format. The transmission spectrum should have a continuous roll-off and the flatness should not exceed -3dB, in order to minimize any signal distortions in the measurement. The gaussmeter probe should then be centered on the fixture to monitor the applied DC
magnetic bias. Stability of the test setup is checked by sweeping the external DC magnetic field from 0Oe to 2kOe, and back to 0Oe. There should be little or no change in the frequency spectrum in response to the magnetic field. The sample is coated with a thin coat of 1827 photoresist and attached to the test fixture. Voltage from the power supply is driven through the amplifier and monitored with the multimeter. With the power supply set to zero volts, the voltage leads are connected to the voltage terminals of the test fixture. Stability of the test setup is checked again by sweeping the external DC magnetic field from 0Oe to 2kOe, and back to 0Oe. Once again, there should be little or no change in the frequency spectrum in response to the magnetic field. The voltage is set to the maximum reverse value (-$V_{\text{max}}$), and the magnetic field is set to the reference value $H_{\text{REF}} \geq 25H_{KDC}$ (where the DC anisotropy field $H_{KDC}$ is determined from VSM measurement). The complex full 2-port S-parameters are measured and stored in a text file. The magnetic field is reduced to the highest bias point to be measured ($H_{\text{bias1}}$), and the full 2-port S-parameters are measured and stored in a text file. The magnetic field is then reduced to next highest bias point ($H_{\text{bias2}}$), and the full 2-port S-parameters are measured and stored in a text file. The measurement sequence is repeated until the S-parameters for all magnetic bias points are measured. In order to capture the hysteretic effect of the electric polarization of the sample, the magnetic bias measurements are repeated for $-0.5V_{\text{max}}$, $-0.25V_{\text{max}}$, $0 V_{\text{max}}$, $0.25V_{\text{max}}$, $0.5V_{\text{max}}$, $V_{\text{max}}$, $0.5V_{\text{max}}$, $0.25V_{\text{max}}$, $0 V_{\text{max}}$, $-0.25V_{\text{max}}$, $-0.5V_{\text{max}}$, and $-V_{\text{max}}$. The FMR spectrum for each bias point is plotted relative to the saturation spectrum.
as, $\text{TransmissionLoss}[dB] = 20 \log_{10} \frac{S_{11 \text{bias}}}{S_{21 \text{ref}}}$. The anisotropy field ($H_K$) is determined from the known sample $4\pi M_S$ by plotting the FMR frequency versus the DC magnetic bias. The FMR linewidth ($\Delta f_{\text{FMR}}$) is determined from the 3dB bandwidth (3dB-BW) of the FMR spectrum. The FMR shift due to electrostatic bias is determined from the change in FMR frequency for each fixed magnetic bias. The ME coupling field ($\Delta H_E$) is derived from the expression $\Delta f_{\text{FMR}} = \gamma \sqrt{\mu_i H_K}$, where $\gamma$ is the gyromagnetic constant 2.8MHz/Oe, and $\mu_i$ is the relative initial permeability.

The permeability spectrum for each magnetic bias point is determined from the relation:

$$\mu_r = \frac{1 + S_{11 \text{bias}} - S_{21 \text{bias}}}{1 - S_{11 \text{bias}}} \cdot \frac{1 + S_{11 \text{ref}} - S_{21 \text{ref}}}{1 - S_{11 \text{ref}}} \cdot \frac{1}{j(k_f \mu_i \omega_o)}.$$

A scaling factor $k_s$ ($0.1 \leq k_f \leq 1$, typically $k_f=0.15$) is determined from curve fitting and extrapolation of the relative permeability back to $\mu_i = \frac{4\pi M_S}{H_K} + 1$ at the bias condition $V_{\text{bias}}=-V_{\text{max}}$, $H_{\text{bias}}=0$Oe, and frequency~0Hz.
Figure 2.5-7. Depiction of FeGaB/PZT composite and test fixture used to measure ME permeability response.
2.6. Electromagnetic (EM) plane wave absorption facility

In order to effectively measure the microwave plane wave absorption of a metamaterial multilayer composite, it is important to know not only the power reflected from the absorber, but also the power transmitted through it as they are interrelated by the expression \( A = 1 - R - T \), where \( R = |S_{11}|^2 \), and \( T = |S_{21}|^2 \). A 2-port EM absorber measurement facility is proposed that consists of an anechoic chamber, a vector network analyzer, a pair of broad band horn antennas, and the necessary rf cabling and interconnecting adapters. Figure 2.6-1 is a depiction of the
anechoic test chamber. The chamber is designed to be easily assembled and disassembled, so that it can be setup and put away as needed (to maximize lab space). The chamber consists of removable panels (Figure 2.6-2) for easy sample loading and to position the horn antennas. The front end panels are bracketed for sliding in the horn antennas. Each end bracket has a back panel that slips behind the antenna closing off the rear of the chamber. The center of the chamber is detachable so that the sample holder can be inserted into place (Figure 2.6-3). This feature provides the option of not using the holder during measurements. Within the sample holder there is a small slot sized just large enough to fit a 4.5x4.5 square inch 30mil thick sample (Figure 2.6-4). There is a 4x4 square inch aperture in the center of the sample holder for which a large portion of the sample is exposed. The aperture was intentionally made smaller than the sample so as not to expose the edges of the metamaterial surface. Figure 2.6-5 and Figure 2.6-6 display the sides of the chamber and its dimensions. The chamber is scaled to satisfy the far-field condition for plane wave radiation for a maximum radiating distance of 4 feet over the frequency range of measurement. The aperture size of the sample holder was set large enough to satisfy both the far field condition and diffraction limit over a narrowband measurement range of 3 to 9 GHz for minimal incident EM plane wave distortion, so that

\[ \sqrt{2ft \times \lambda_{\text{max}} \times \frac{1}{2}} \geq D_{\text{aperture}} > \lambda_{\text{max}} \], where \( \lambda_{\text{max}} \) is the maximum guided wavelength.

The measurement range can be adjusted based on the aperture size of the sample holder. The full measurement range of the chamber would be ~1GHz to 12GHz based on the bandwidth of the horn antennas. The chamber housing would be
constructed of a reinforced wood framework and held together with steel brackets (commonly used as shipping crates) for structural soundness and ease of disassembly. The chamber is also sized to accommodate 20 inch radar absorbing pyramid foam panels which would line the walls of the inner chamber, and 4 inch absorbing panels on both sides of the sample holder (Figure 2.6-7). The radar absorbing panels are intended to attenuate background noise during measurement over a broad frequency band. The chamber must be made sufficiently quite so that the noise floor (empty chamber condition) is well below that of the reflected and transmitted signals measured from the test sample. A noise floor below 40dB is desired, to minimize backscattering noise that would introduce errors in the measurement.

The measurement procedure is would be performed as follows: (1) measure $R_{\text{noise}}$ and $T_{\text{noise}}$ (noise floor measurement), (2) measure $R_{\text{open}}$ and $T_{\text{open}}$ (empty sample holder measurement), (3) measure $R_{\text{short}}$ and $T_{\text{short}}$ (copper reflector sample measurement), and (4) measure $R_{\text{sample}}$ and $T_{\text{sample}}$ (metamaterial sample measurement). The purpose of (1) is to ensure that the chamber is sufficiently quite (<40dB noise floor). The measurement of steps (2) and (3) are used to normalize the transmission and reflection coefficients respectively of the measured results of step (4) [2.9]. Time domain gating [2.10] would be used to isolate wave scattering from the sample holder and tested sample from the background noise.

Metamaterial normal incident EM plane wave absorber designs were simulated using HFSS10/HFSS11, by simulating an infinite two dimensional lattice of unit cells from a single unit cell with master and slave periodic boundaries [2.11].
Perfectly matched layers (PMLs) were set up at the top and bottom ends of an air box as shown in Figure 2.6-8 to minimize EM wave backscattering. An EM plane wave emanates from the top PML towards the bottom PML. The bottom PML is configured to function as the reference plane (using the FSS reference feature in HFSS), so that the reflection ($S_{11}$) and transmission ($S_{21}$) coefficients can be determined. The simulation uses two air boxes. The dimensions of the inner air box is $p \times p \times \left( \frac{\lambda_{\text{max}}}{2} + t_{\text{substrate}} + 2 \cdot t_{\text{cladding}} \right)$, where $p$ is the unit cell size and lattice spacing, $\lambda_{\text{max}}$ is the maximum guided wavelength incident on the test sample, and $t_{\text{substrate}} + 2 \cdot t_{\text{cladding}}$ is the thickness of the multilayer test sample (which includes the substrate thickness and combined thickness of the top and bottom conducting elements). This provides sufficient radiating distance to ensure an adequate noise floor as shown in Figure 2.6-9. An outer air box encloses both the inner air box and the two PML boxes, and has the dimension $p \times p \times \left( h_{\text{inner-\:airbox}} + 2 \cdot h_{\text{PML}} \right)$, where $h_{\text{PML}} = \frac{\lambda_{\text{max}}}{12}$ is the height of the PML box. The outer air box was used to set up continuous 2-D master-slave boundary pairs along the sides of the PML and inner air box. The master-slave boundaries are configured for plane waves normally incident on the surface of the test sample. The simulation frequency is swept from 1 to 10GHz, typically in steps of 0.1 GHz.
Figure 2.6-1. Proposed EM wave absorption measurement chamber construction.

Figure 2.6-2. Panels utilized during routine setup and measurement.
Figure 2.6-3. Description of chamber sections.

Figure 2.6-4. Sample holder construction.
Figure 2.6-5. Side view of test chamber, depicting approximate radiating distance.

Figure 2.6-6. Dimension of each chamber section.
Figure 2.6-7. Sidewalls lined with 20inch radar absorbing foam.
Figure 2.6-8. HFSS Simulation. Set-up of PML boundaries.

**Effective Noise Floor (No Sample)**

Figure 2.6-9. Measurement of empty chamber condition to test PML boundaries.
2.7. References

Chapter 3. Coupling phenomena in layered composites

Layered composites are formed by depositing alternating layers of materials with different properties. These alternating multilayered stacks are periodic in nature so that the interactions between layers produce composite structures that exhibit properties not found in the individual layers. [3.1] This section provides an overview of the theories of coupling phenomena in exchange biased and magnetoelectric multilayers utilized in this paper.

3.1. Exchange coupling in FM/AFM/FM multilayers

One of key motivations for using metallic magnetic multilayers is that the magnetization can be increased, while minimizing conductive losses in the magnetic layers. The effective magnetization may then be expressed as

\[ M_{\text{eff}} = \frac{\sum t_{\text{FM}} M_s}{\sum t_{\text{FM}} + \sum t_{\text{NM}}} \]

where \( t_{\text{FM}} \) and \( t_{\text{NM}} \) are the magnetic layer and nonmagnetic spacer layer thicknesses. If the nonmagnetic layers are replaced with antiferromagnetic layers, the result is a uniaxial anisotropy field known as the exchange bias field (\( H_{\text{ex}} \)), which results in an enhanced anisotropy field

\[ H_{k,\text{eff}} = H_{k,i} + H_{\text{ex}} + \Delta H_c \]

where \( H_{k,i} \) is the intrinsic anisotropy field, and \( \Delta H_c \) is an increase in the hard axis coercivity due exchange coupling. This in turn results in an increase in the FMR frequency as expressed by

\[ f_{\text{FMR}} = \gamma \cdot \sqrt{(H_{k,\text{eff}} + H_{\text{FMR}}) \cdot 4\pi M_s} \]

where \( H_{\text{FMR}} \) is the FMR field, and \( \gamma = 2.8\text{MHz/Oe} \) is the gyromagnetic ratio.
However, the FMR linewidth ($\Delta H_{\text{FMR}}$), is also increased in proportion to $\Delta H_{c}$. $\Delta H_{c}$ is believed to be due to uncompensated magnetic charges at the edges of the multilayer structure [3.2], where the magnetic charge density ($\vec{\nabla} \cdot \vec{M}$) may be expressed in terms of the demagnetizing field ($\vec{H}_{D}$) as $-4\pi\vec{\nabla} \cdot \vec{M} = \vec{\nabla} \cdot \vec{H}_{D}$ [3.1].

By sandwiching the antiferromagnetic layer with two magnetic layers, the magnetic charges are better compensated, resulting in lower $\Delta H_{c}$.

The exchange bias field is inversely related to the magnetic layer thickness by $H_{\alpha} = \frac{J_{\alpha}}{M_{s}J_{\text{FM}}}$, (where $J_{\alpha}$ is the interfacial exchange energy). Thus, minimizing the thickness of the magnetic layers in the multilayer composite increases the FMR frequency.

### 3.2. Magnetoelectric coupling in ME layered composites

A layered ME composite consists of at least one piezoelectric layer known as the piezoelectric phase, and a magnetic layer called the magnetic phase. Elastic coupling between these two phases leads to the tuning of dielectric polarization under an applied magnetic field or the control of an induced magnetization under an external electric field, thus enabling an effective energy conversion between electric and magnetic fields (Figure 3.2-1). This magnetoelectric response is known as the magnetoelectric (ME) effect.
When a transverse voltage $V$ is applied across the thickness $t$ of the piezoelectric phase, a biaxial strain of $\Delta l/l = d_{31}(V/t)$ is induced, where $d_{31}$ is the piezoelectric charge coefficient. This phenomenon is known as the inverse piezoelectric effect. Deformation of the piezoelectric phase will lead to strain in the magnetic phase which is tightly bonded to the piezoelectric phase. This induced strain leads to a change in the in-plane anisotropy field due to the well known magnetoelastic effect, which is reflected in a shift in the FMR frequency of the magnetic film at microwave frequencies as described by the relation $\Delta f_{\text{FMR}} = \gamma \sqrt{\mu_i \Delta H_{\text{eff}}}$, where $\mu_i$ is the intrinsic permeability, and $\Delta H_{\text{eff}}$ is the effective in-plane magnetization field.

For metallic magnetic thin films $\Delta H_{\text{eff}} = 3\lambda_s Y d_{31} \times \frac{E}{M_s}$ (in CGS units), where $\lambda_s$, $M_s$, and $Y$ is the saturation magnetostriction, Young’s modulus, and saturation magnetization of the magnetic phase respectively, while $E$ is the transverse electrostatic field applied across the piezoelectric phase. The larger the $\lambda_s$ and $d_{31}$ of the magnetic phase and piezoelectric phase respectively the larger the FMR.
shift $\Delta f_{FMR}$, for a given applied field $E$. The quality of the bond between the magnetic and piezoelectric layer may constrain the coupling strength between phases, so that $\Delta H_{\text{eff}} \leq 3\lambda_s y d_{31} \times \frac{E}{M_s}$. For YIG films where the crystal plane normal is $[111]$ and in-plane magnetization is along $[1\overline{1}0]$, the effective in-plane magnetization field is similarly expressed as $\Delta H_{\text{eff}} \leq -\frac{3}{2} (\lambda_{100} + \lambda_{111}) y d_{31} \times \frac{E}{M_s}$.

3.3. References


Chapter 4. DC measurement results

4.1. Ferromagnetic/antiferromagnetic/ferromagnetic trilayer characterization

The hysteresis loops along the easy axis (EA) show clear hysteresis loop shift due to exchange coupling from the AFM layer, while the hard axis (HA) hysteresis loops are typically slim with no hysteresis shift, as shown in Figure 4.1-1. The EA coercivity ($H_{c_{-EA}}$), HA coercivity field ($H_{c_{-HA}}$), effective anisotropy field ($H_{k_{-HA}}$), exchange bias field along the easy axis ($H_{ex}$), and $H_{ex}+H_{c_{-EA}}$ were plotted versus the magnetic layer thickness $t_{FM}$ for each of the sandwich thin-film sample sets of Co$_{90}$Fe$_{10}$[Ru], Co$_{84}$Fe$_{16}$[Ru], and Co$_{84}$Fe$_{16}$[NiFeCr] in Figure 4.1-2. The effective anisotropy field $H_{k_{-HA}}$ decreases with the increase of $t_{FM}$, and matches well with the $H_{ex}+H_{c_{-EA}}$ curve in all three sample sets, indicating that $H_{k_{-HA}} \approx H_{ex}+H_{c_{-EA}}$, as observed in different thin-film systems. [4.1] The two sets of sandwich films Co$_{90}$Fe$_{10}$[Ru] and Co$_{84}$Fe$_{16}$[Ru] both show a low hard axis coercivity of 2–4 Oe, a squareness ratio of >98%, and a relatively large $H_{ex}$ of over 40 Oe at $t_{FM} = 100$ Å. Both the low hard-axis coercivity and enhanced anisotropy field are desired for rf/microwave applications.
Figure 4.1-1. In-plane easy-axis and hard-axis hysteresis of a NiFeCr seeded CoFe/PtMn/CoFe sample.
Figure 4.1-2. Easy axis coercivity ($H_{c\_EA}$), hard-axis coercivity field ($H_{c\_HA}$), effective anisotropy field ($H_{k\_HA}$), and exchange bias field along easy-axis ($Hex$), versus the magnetic layer thickness ($t_F$) of (top) $\text{Co}_{90}\text{Fe}_{10}/\text{Pt}_{50}\text{Mn}_{50}/\text{Co}_{90}\text{Fe}_{10}$ with Ru, (center) $\text{Co}_{84}\text{Fe}_{16}/\text{Pt}_{50}\text{Mn}_{50}/\text{Co}_{84}\text{Fe}_{16}$ with Ru and (bottom) $\text{Co}_{84}\text{Fe}_{16}/\text{Pt}_{50}\text{Mn}_{50}/\text{Co}_{84}\text{Fe}_{16}$ with NiFeCr, sandwich thin-film sample sets.

The sandwich film set $\text{Co}_{84}\text{Fe}_{16}[\text{NiFeCr}]$, however, shows higher hard axis coercivities in the range of 7–25 Oe, and a very low $H_{ex}$ less than 7 Oe, when $t_{FM}$ is
in the range of 100–500 Å. The interfacial exchange energies can be calculated to be about 0.058, 0.049, and 0.023 erg/cm$^2$ for the three sets of samples Co$_{90}$Fe$_{10}$[Ru], Co$_{84}$Fe$_{16}$[Ru], and Co$_{84}$Fe$_{16}$[NiFeCr], respectively. The effective anisotropy fields of the CoFe films was enhanced to be ~160 Oe for both Co$_{90}$Fe$_{10}$[Ru] and Co$_{84}$Fe$_{16}$[Ru], corresponding to a boosted FMR frequency of above 5 GHz at zero bias field.

The low hard axis coercivity and relatively large exchange bias field of the two sets of sandwich films Co$_{90}$Fe$_{10}$[Ru] and Co$_{84}$Fe$_{16}$[Ru] are associated with a fine grain size for the PtMn and CoFe layers, as indicated by the broad and low-intensity PtMn {111} and CoFe {110} x-ray diffraction peaks in the out-of-plane “$\theta - 2\theta$” XRD patterns in Figure 4.1-3. The Co$_{84}$Fe$_{16}$[NiFeCr] film set shows significantly higher intensity and narrower PtMn {111} and CoFe {110} diffraction peaks. Grain-size evaluation using the Scherrer equation [4.2] was done with the PtMn {111} and FeCo {110} peaks of the XRD pattern, as indicated in Table 4.1-1. The Co$_{84}$Fe$_{16}$[NiFeCr] film shows a relatively large grain size of 17 nm for the PtMn and 31 nm for the CoFe phase than those of the Co$_{90}$Fe$_{10}$[Ru] and Co$_{84}$Fe$_{16}$[Ru], which show similar grain size of 12 nm for the PtMn and 16 nm for the CoFe phase. The large grain sizes in the PtMn and CoFe layers in the Co$_{84}$Fe$_{16}$[NiFeCr] film sets may account for their high hard axis coercivity and low exchange bias fields.
Figure 4.1-3. XRD results for the samples in the sample sets of Co$_{90}$Fe$_{10}$[Ru], Co$_{84}$Fe$_{16}$[Ru], and Co$_{84}$Fe$_{16}$[NiFeCr] with a CoFe layer thickness of 200 Å.

Table 4.1-1. Grain size obtained from the Scherrer equation with the PtMn {111} and CoFe {110} peaks. Uncertainty of the grain-size evaluation is ~10%.

<table>
<thead>
<tr>
<th>Sample ID</th>
<th>PtMn grain size (nm)</th>
<th>CoFe grain size (nm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Co$<em>{90}$Fe$</em>{10}$[Ru]</td>
<td>12</td>
<td>16</td>
</tr>
<tr>
<td>Co$<em>{84}$Fe$</em>{16}$[Ru]</td>
<td>12</td>
<td>16</td>
</tr>
<tr>
<td>Co$<em>{84}$Fe$</em>{16}$[NiFeCr]</td>
<td>17</td>
<td>31</td>
</tr>
</tbody>
</table>

4.2. Magnetoelectric composite characterization

The magnetic hysteresis of the YIG/PZT bilayer was measured on a vibrating sample magnetometer. Low-frequency ME voltage coupling coefficient of the YIG/PZT bilayer composite was measured under a sinusoidal magnetic field...
excitation at 85 kHz under different bias fields. The results for both are shown in Figure 4.2-1. Clearly, the ME voltage coefficients peak at +10 and -10 Oe and disappear beyond 50 Oe, which correlates well with the hysteresis loop of the YIG material, which shows saturation at fields above 50 Oe. This correlation indicates that the low-frequency ME response from magnetic field excitation originates from the magnetization process, mainly due to domain activities such as wall motion and domain rotation of the YIG layer, consistent with what was reported [4.3-4.5]. The dc and rf characteristics of the metallic magnetic films used in this study are described in Table 4.2-1.

**Low Frequency Response**

![Figure 4.2-1. Continuous wave response of ME voltage coefficient (in kilovolts per centimeter per oersted) versus dc magnetic field (in oersted). Axis 2: In-plane hysteresis of M/Ms.](image)
Table 4.2-1. Magnetic and microwave properties of the ME metallic magnetic films.

<table>
<thead>
<tr>
<th>Composition [at %]</th>
<th>$4\pi M_s$ ($\text{Oe}$)</th>
<th>$H_c$ ($\text{Oe}$)</th>
<th>$H_k$ ($\text{Oe}$)</th>
<th>$\Delta H$ at x-band ($\text{Oe}$)</th>
<th>$\lambda_s$ (ppm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fe$<em>{75}$Co$</em>{15}$B$_{10}$</td>
<td>17</td>
<td>&lt;1</td>
<td>19</td>
<td>&lt;20</td>
<td>40</td>
</tr>
<tr>
<td>Fe$<em>{72}$Ga$</em>{10}$B$_{18}$</td>
<td>12</td>
<td>&lt;2</td>
<td>21</td>
<td>15</td>
<td>50-70</td>
</tr>
</tbody>
</table>

4.3. References

4.2. A. Guinier, X-ray Diffraction (Freeman, San Francisco, 1963), p. 121.
Chapter 5. RF measurement results

5.1. FM/AFM/FM trilayer results and analysis

5.1.1. In-Plane FMR behaviour of single period trilayers at X-band

The in-plane FMR linewidth parallel to the exchange bias field axis \( H_{ex} \) (0°) of the Ru and NiFeCr seeded CoFe/PtMn/CoFe samples are shown in Figure 5.1-1 as a function of the CoFe FM layer thickness \( t_{FM} \). The error bar for the FMR linewidth of the trilayer samples with a \( t_{FM} \) of 50 Å and 100 Å is in the range of 10–20 Oe, which drops to about 3–5 Oe for the trilayer samples with thicker \( t_{FM} \) due to the relatively sharper FMR absorption peaks. The Co\(_{84}\)Fe\(_{16}\)[NiFeCr] sample set shows a monotonically decreasing FMR linewidth with the increase of the \( t_{FM} \), with the minimum FMR linewidth of 100 Oe. This behavior is similar to what is observed in the exchange-biased NiFe/NiO bilayer films.[5.1, 5.2] The Co\(_{90}\)Fe\(_{10}\)[Ru] and Co\(_{84}\)Fe\(_{16}\)[Ru] sample sets, however, show different behavior, wherein the minimum FMR linewidth is obtained at an intermediate \( t_{FM} \) of 200 Å, rather than at the maximum \( t_{FM} \). This behaviour differs from what was reported for the FM/AFM bilayers, which typically show a monotonic drop with the increment of the FM layer thickness [5.1-5.9]. The minimum FMR linewidth among all three sample sets was achieved for the 200 Å thick Co\(_{84}\)Fe\(_{16}\)[Ru] film, which is 45 Oe. The relatively large FMR linewidth of the Co\(_{84}\)Fe\(_{16}\)[Ru] sample set at 400 and 500 Å CoFe layer
thicknesses is related to the appearance of an anomalous absorption peak that interferes with the uniform mode FMR peak, as shown in Figure 5.1-2, which is in the Co$_{84}$Fe$_{16}$[Ru] sample set at 100–500 Å. The anomalous absorption peak, however, is not observed in either the Co$_{90}$Fe$_{10}$[Ru] or Co$_{84}$Fe$_{16}$[NiFeCr] sample set. The FMR linewidth in Co$_{90}$Fe$_{10}$[Ru] increases with the increase of the $t_{FM}$ when the $t_{FM}$ is over 200 Å, which cannot be explained by the enhanced eddy current damping with the increased film thickness, as the total film thickness in all film sets is less than or equal to 1120 Å, which is significantly lower than the skin depth at x-band, which is calculated to be about 4000 Å. The Co$_{84}$Fe$_{16}$[Ru] and Co$_{90}$Fe$_{10}$[Ru] sample sets were further examined to investigate this phenomena, by examining the in-plane rotational characteristics.
Figure 5.1-1. In-plane X band (~9.5 GHz) FMR Linewidth ($\Delta H$) versus magnetic CoFe layer thickness $t_{FM}$ of the samples of Co$_{90}$Fe$_{10}$[Ru], Co$_{84}$Fe$_{16}$[Ru], and Co$_{84}$Fe$_{16}$[NiFeCr].

Figure 5.1-2. Field sweep FMR spectrum showing the differential absorption power with respect to field vs the applied in-plane dc magnetic field for the Co$_{84}$Fe$_{16}$[Ru] with a CoFe layer thickness of 200 Å.
Figure 5.1-3 is a plot of the difference in FMR field \((H_{\text{FMR}} -180^\circ) - (H_{\text{FMR}} -0^\circ)\) parallel \((0^\circ)\) and antiparallel \((180^\circ)\) to the \(H_{\alpha}\) direction as a function of \(t_{\text{FM}}\). The error bar for these measurements (not shown in the Figure) is 2–3 Oe. This FMR field difference is expected to be twice \(H_{\alpha}\) at FMR frequency. The extracted FMR \(H_{\alpha}\) field versus the \(t_{\text{FM}}\) of the Co\(_{90}\)Fe\(_{10}\)[Ru] trilayer roughly correlates with the dc \(H_{\alpha}\) (Figure 4.1-2) but higher in value by as much as 75\%. On the contrary, the Co\(_{84}\)Fe\(_{16}\)[Ru] trilayer FMR \(H_{\alpha}\) does not correlate with the dc results and trend lower by as much as 98\%. In examining the profile for the Co\(_{90}\)Fe\(_{10}\)[Ru] sample set, the 200 Å structure shows a clear presence of exchange bias field at FMR frequency as indicated by the difference between the 0\(^\circ\) and 180\(^\circ\) results, while no exchange bias field is observed for the 300 Å structure. To confirm this odd behavior, a full rotational \((0^\circ – 360^\circ\) away from the dc exchange bias field direction\) FMR field \((H_{\text{FMR}})\) measurement was taken for these two structures. The plotted \(H_{\text{FMR}}\) versus rotation shown in Figure 5.1-4 produce butterfly-shaped patterns that give an indication of exchange bias. The symmetrical pattern in the 300 Å trilayer indicated that \(H_{\alpha}\) at FMR frequency is roughly zero, consistent with Figure 5.1-3. This observed behavior appears to be linked to the anomalous absorption peak (or mode), shown in Figure 5.1-5, which is close to the uniform mode when the CoFe layer thickness is 300 Å or above. The anomalous mode is at the left side of the uniform mode at \(t_{\text{FM}} = 300\) Å, which shifts to the right side of the uniform mode when \(t_{\text{FM}}\) increases. The origin of this anomalous mode is not clear at this time. However, this
behavior, once fully understood, could conceivably be utilized during thin-film growth to optimize the linewidth.

Figure 5.1-3. Difference in FMR field with the applied field at 0° and 180° with respect to the CoFe layer thickness.
Figure 5.1-4. Plot of $H_{FMR}$ versus in-plane rotation for Co$_{84}$Fe$_{16}$/Pt$_{50}$Mn$_{50}$/Co$_{84}$Fe$_{16}$ with Ru seed and cap layer for CoFe layer thicknesses of 200 and 300 Å.

Figure 5.1-5. FMR versus in-plane bias field along the 0° axis for Co$_{84}$Fe$_{16}$/Pt$_{50}$Mn$_{50}$/Co$_{84}$Fe$_{16}$ with Ru seed and cap layer with different CoFe layer thicknesses as indicated in the chart.
The FMR fields for both sample sets are shown in Figure 5.1-6. The FMR field increases monotonically, with the increase in $t_{FM}$, from 220 Oe to as high as 550 Oe. According to the Kittel equation $f_{FMR} = \gamma \cdot \sqrt{(H_{K,eff} + H_{FMR}) \cdot 4\pi M_s}$ with $\gamma \cong 2.8$ MHz/Oe, $f_{FMR} = 9.6$ GHz and $4\pi M_s = 20$ kG for the CoFe alloys, we get $H_{K,eff} + H_{FMR} = 560$ Oe. This indicates an effective anisotropy field $H_{K,eff}$ in the range of 10–340 Oe for the CoFe/PtMn/CoFe trilayers, corresponding to a self-biased FMR frequency in the range of 1.8–7.2 GHz. The combination of low FMR linewidth, high saturation magnetization, the low coercivity, and significantly enhanced effective magnetic field in the Ru seeded CoFe/PtMn/CoFe trilayer films make them great candidates for rf/microwave frequency applications.

![Figure 5.1-6. FMR field versus magnetic CoFe layer thickness $t_{FM}$ for the two sets of trilayers as indicated in the figure.](image-url)
It is notable that the effective anisotropy fields of these CoFe/PtMn/CoFe trilayer films at microwave frequency are completely different from those at dc frequency. At a CoFe layer thickness of 100 Å, the anisotropy fields of the trilayers at x-band are more than twice those at dc frequency. Rodriguez–Suarez et al. [5.10] attribute this discrepancy to the magnetic behavior of stable and unstable AFM grains at the AFM–FM interface. The conditions of the FMR measurements are different because the static applied field values are much larger than the applied fields in the dc VSM measurements. However, they observe $H_{ex}$ and $H_K$ values at FMR frequency to be consistently lower than at dc, contrary to our observations. Further investigation is needed to understand why the FMR linewidth in Co$_{90}$Fe$_{10}$[Ru] increases with the increase of the FeCo layer thickness.

The effective anisotropy fields, coercivities, and FMR linewidth of the eight-period structures multilayers are shown in Table 5.1-1. The eight-period multilayer based on Co$_{84}$Fe$_{16}$[Ru] trilayer has the best combination of magnetic and microwave properties, a high anisotropy field of 121 Oe, a low hard axis coercivity of 3.5 Oe, an easy axis coercivity of 66 Oe, and a FMR linewidth of 132 Oe. Further work is needed to understand the origin of this difference between the effective anisotropy fields at dc and microwave frequencies, in order to fully exploit this behavior for microwave applications.
Table 5.1-1. Magnetic and microwave properties of the 8-period multilayers.

<table>
<thead>
<tr>
<th>8-period film structures</th>
<th>anisotropy field (Oe)</th>
<th>Hard axis coercivity (Oe)</th>
<th>Easy axis coercivity (Oe)</th>
<th>FMR linewidth at 9.5 GHz (Oe)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Al₂O₃/Co₉₀Fe₁₀[Ru]</td>
<td>121</td>
<td>4.5</td>
<td>72</td>
<td>148</td>
</tr>
<tr>
<td>Al₂O₃/Co₈₄Fe₁₆[Ru]</td>
<td>121</td>
<td>3.5</td>
<td>66</td>
<td>132</td>
</tr>
</tbody>
</table>

5.2. ME composite results and analysis

5.2.1. Dynamic ME behaviour in ferrite/piezoelectric (PZT) composites

The FMR frequency of the YIG/PZT bilayer was measured at different bias magnetic fields and with zero applied electric field, which can be well fitted to the Kittel equation with \( H_K = 42 \text{ Oe}, \) and \( 4\pi M_S \) of 1950 G for magnetic thin film with in-plane magnetization. \( f_{\text{FMR}} = \gamma \sqrt{(H_K + H_{\text{DC}})(4\pi M_S + H_K + H_{\text{DC}})} \), where \( \gamma \) is about 2.8 MHz/Oe. The result is plotted in Figure 5.2-1. The ME coupling at microwave frequencies was then investigated over a broad frequency range of 1–7 GHz under different magnetic and electric bias fields. The YIG/PZT bilayer was loaded into the air gap of a microstrip, which generates a broadband excitation as described in section 2.5.1. The FMR response was measured as a function of both an applied dc magnetic field and an applied dc electric field. The dc magnetic field was swept from +1600 to -1600 Oe. At each magnetic field bias, the FMR response was measured along the range \( \pm 7.2 \text{ kV/cm} \). The FMR spectrum at 1kOe, with the electric field swept from +7.2kV/cm to -7.2kV/cm is shown in Figure 5.2-2. In this experiment, the YIG layer
has a plane normal of [111]. Assuming both the stress and the magnetization of the YIG layer is in-plane along the [110] direction, the magnetoelastic energy will be

\[ E_{ME} = -\frac{1}{2} \sigma (\lambda_{100} + \lambda_{111}) \]

leading to a stress-induced effective magnetic field strength

\[ \delta H_E = -3(\lambda_{100} + \lambda_{111})\sigma / 2M_S \equiv A \cdot E \] (cgs unit), where \( A \) is defined as the ME constant and \( E \) the electric field applied on the ME bilayer. Depending on the tensile or compressive stress, the stress-induced effective magnetic field \( \delta H_E \) will change sign, leading to a shift of the FMR frequency of

\[ \delta f_{FMR} = \gamma^2 [4\pi M_s + 2(H_K + H_{DC} + \delta H_E)] \cdot \delta H_E / f_{FMR}. \]

The YIG/PZT device showed 20–30 MHz FMR frequency shift at a bias magnetic field of 1000–1500 Oe with electric field biased in the range ±7.2 kV/cm. Plugging in the electric field of 7.2 kV/cm, \( \lambda_{100} = -1.4 \) ppm, and \( \lambda_{111} = -2.85 \) ppm [5.11], we get \( \delta H_E \) in the range of 7–10 Oe, \( A \) in the range of 0.5–1 Oe cm/kV, and a stress \( \sigma \) exerted on the YIG of 200–300 Mdyne/cm (or 20–30MPa). The magnetoelectric constant of 0.5–1 Oe cm/kV observed in this paper is comparable to what was predicted [5.12]. According to the above equation, the FMR frequency shift will have a strong dependence on the applied field, particularly at small bias fields when the FMR frequency is low. Interestingly, this field dependence is not observed in our experiments, which may be due to the complication of the magnetostatic spin waves.
Figure 5.2-1. FMR frequency of the YIG/PZT bilayer measured at different bias magnetic fields and with zero applied electric field.

Figure 5.2-2. FMR shift at magnetic bias field of 1kOe, with the electric field swept from +7.2kV/cm to -7.2kV/cm.
5.2.2. Dynamic ME behaviour in metal magnetic thin film/PZT composites

A broad band coplanar permeability measurement technique was used to measure the permeability of the metallic magnetic thin film/PZT composites, in response to both an in-plane magnetostatic field along the easy axis direction and a transverse electrostatic bias field across the thickness of the PZT slab. This technique is described in section 2.5.2. An electrostatic bias sweep in the range ±8 kV/cm was applied to the PZT layer by biasing the electrodes at different voltages between −400 V and +400 V.

The measured permeability spectra of the FeCoB/PZT composite film at different applied voltages across the PZT layer is shown in Figure 5.2-3 under a constant bias field of 20 Oe. An FMR frequency shift of 50 MHz was obtained at 2.3 GHz together with a negligibly small change of the initial permeability. The FeGaB composite was biased at 35 Oe to observe the ME effect at the same frequency. The permeability spectra of the FeGaB/PZT ME composite are shown in Figure 5.2-4 at different applied voltages across the PZT layer. The measured FMR frequency shift in the range ±8 kV/cm for the FeGaB/PZT composite is ~110 MHz, which is much larger than the 50 MHz FMR frequency shift for the FeCoB/PZT ME composites.
Figure 5.2-3. Permeability spectrum of FeCoB/PZT composite vs electrostatic bias at a fixed magnetostatic bias of 20 Oe (~2.3 GHz).

Figure 5.2-4. Permeability spectrum of FeGaB/PZT composite vs electrostatic bias at a fixed magnetostatic bias of 35 Oe (~2.3 GHz).
In both the FeCoB/PZT and FeGaB/PZT ME composites, the application of the negative electric field causes a downward shift in the permeability spectrum, while a positive field produces the opposite response. This behavior could be explained by considering the electric field induced the elastic deformation of the PZT and its effect on the anisotropy field. A tensile strain in the metallic magnetic films with a positive magnetostriction constant leads to an increase in the effective in-plane anisotropy field, whereas a compressive strain leads to a decrease. The effective in-plane magnetic field ($\Delta H_{\text{eff}}$) in a magnetic film/ferroelectric ME composite bilayer generated from the stress mediated ME coupling can be expressed as follows (cgs unit): $\Delta H_{\text{eff}} = 2\lambda_d Y d_{31} \times \frac{E}{M_S}$, where $M_S$ is the saturation magnetization, and $Y$ is the Young’s modulus of the magnetic film; $d_{31}$ is the piezoelectric coefficient and $E$ is the electric field across the thickness direction of the ferroelectric substrate. Since the FMR phenomenon is used for many microwave magnetic devices,[5.3, 5.13, 5.14] the tunable range of the FMR frequency is a representation of the tunability of many kinds of such microwave ME devices. The tunable FMR frequency range ($\Delta f_{\text{FMR}}$) induced by the effective in-plane magnetic field ($\Delta H_{\text{eff}}$) of such microwave ME devices, with metallic magnetic film deposited directly on a ferroelectric substrate, can be derived to be $\Delta f_{\text{FMR}} = \gamma \sqrt{\mu_i \Delta H_{\text{eff}}}$, where $\gamma$ is the gyromagnetic constant ~2.8 MHz/Oe, and $\mu_i$ is the initial relative permeability of the magnetic film. We can readily get $\Delta f_{\text{FMR}} / f_{\text{FMR}} = \Delta H_{\text{eff}} / H_{DC}$, where $H_{DC}$ is the net in-plane uniaxial anisotropy field of the magnetic film, and $f_{\text{FMR}}$ being the FMR frequency, or operating...
frequency in many different microwave devices.[5.3, 5.13, 5.14] For the microwave ME composite with Fe$_{75}$Ga$_{13}$B$_{12}$ (at %) film on a PZT slab, we will get a $\Delta H_{\text{eff}}$ of ~30 Oe and a large tunable FMR frequency range of $\Delta f_{\text{FMR}} = 650$ MHz at 2.5 GHz when a moderate electric field of 800 V/mm is applied across the PZT layer by the predicted FMR peak shift (Figure 5.2-5) derived from the Landau–Lifshitz–Gilbert equation.[5.15] This corresponds to a large tunability of $\Delta f_{\text{FMR}} / f_{\text{FMR}} = 28\%$. This large tunability in the FeGaB/PZT bilayer ME composite material is one order of magnitude higher than what was reported for the ME composites, which are typically in the range of $\Delta f_{\text{FMR}} / f_{\text{FMR}} = 0.5 – 2.5\%$ with $\Delta f_{\text{FMR}} = 30–125$ MHz.[5.14, 5.16, 5.17]

For the ME composites reported, there is always a substrate for the metallic magnetic film, i.e., FeCoB/glass/PZT and FeGaB/silicon/PZT. The presence of a substrate for the metallic magnetic films as well as the non-ideal strain coupling at the interface of the ME composite may explain the discrepancy between the observed and predicted performances. Putting metallic magnetic films onto PZT directly is hindered by the large magnetic linewidth of the metallic magnetic films, which requires further investigation.
Figure 5.2-5. Simulated real (blue) and imaginary (red) permeability spectra of a FeGaB/PZT ME composite at two E-fields (0: solid line; and 800V/mm: dot-dash line) under a constant bias magnetic field of 20 Oe.

5.3. References

5.13. J. D. Adam, L. E. Davis, G. F. Dionne, E. F. Schloemann, and S. N.
Chapter 6. Microwave applications and challenges

6.1. Microwave integrated filters

6.1.1. Motivation and challenges

ME based tunable filters have been actively researched as an alternative to current driven tunable filters which require an electromagnet to tune the FMR response of the magnetic material for band filtering. The current driven filter has the advantage of wide tunable range, limited primarily by the induced field capability of the internal electromagnet. However, the power consumption required for driving the electromagnet, and its relative scale makes it impractical as a MMIC filter. An ME based filter has the benefit of providing two means of tuning the filter response. It can be driven magnetostatically using a current source, and electrostatically using a voltage source. However, its greater benefit is the voltage tunability because it eliminates need for an electromagnet for tunability. The tunability is driven mainly by the properties of the ME composite. MMIC devices are often voltage driven due to less power consumption. This is especially important for mobile applications. So a voltage driven tunable filter would appear to be easier to integrate into a MMIC chip. However, there are still many issues to resolve with ME composites. First, ME composites are essentially electromechanically driven. A mechanical stress is induced using an electrostatic bias. This implies that the composite must be moving
in response to the induced stress. MMIC devices are essentially solid state, and therefore have no moving parts. Second, ME composites to date still require a relatively high voltage to achieve a dramatic tunable range [6.1, 6.2]. The voltage range needs to be in the order of single digit volts as opposed to hundreds of volts to make them practical to use at the MMIC scale. Third, the methods for synthesizing ME composites must be compatible to MMIC fabrication. This implies that the layered composites must be formed on the substrate along with the rest of the circuit using a low temperature process. Fourth, there is still the question of the reliability of ME composites [6.1]. How much stress can it take over time before failure (eg. stress fractures in the piezoelectric or magnetic layers). This section will provide an overview of the main criteria for high performance ME based tunable filters, and explore efforts to resolve some of the issues associated with ME composites.

The key features for a high performance ME tunable filter are narrow bandwidth for high selectivity, strong FMR absorption for high attenuation, high magnetostriction for wide tunability, strong ME coupling constant for low voltage biasing, and high FMR frequency for microwave operation. The selectivity of a filter whether it be band-reject or band-pass is a measure of how narrow the filtering bandwidth is relative to the total operating bandwidth. The higher the selectivity, the higher the filter resolution is. In other words, the better a filter is able to discriminate between one frequency and another. For applications requiring communication channel selection, high selectivity allows for more frequency channels to be packed into the total operating bandwidth. Selectivity may be defined as the slope of the
magnitude response of the filter at the normalized corner frequency, or band edge (in other words a measure of cut-off rate) \( -\frac{d|H(j\omega)|}{d\omega} \) where \( \omega \equiv 1 \) [6.3]. Attenuation is a measure of the filters ability to isolate the filtered bandwidth signal from the out of band signals. The higher the attenuation the lower the interaction with out of band signals. Attenuation may be defined as \( 10\log\left(\frac{P_{\text{out}}}{P_{\text{in}}}\right)\) dB [6.4]. The tunability of a filter is a measure of its ability to shift the center frequency of the filtering bandwidth anywhere within the total operating bandwidth. Tunability may be defined as \( \frac{\Delta f}{f_o} \times 100\% \), where \( f_o \) is the operating frequency, and \( \Delta f \) is the tunable bandwidth.

The tunable filter is an active device, and therefore requires an external bias to tune the center frequency. To reduce the power consumption of the device it is desirable for the external bias (whether it be voltage or current) to be as low as possible. So the more responsive the center frequency is to changes in the voltage the less voltage needed to tune the filter to the target frequency. Therefore the ME coupling coefficient [6.5], is an important figure of merit for filter response. The maximum operating bandwidth depends not only on the tunable range of the filter, but also the unbiased FMR frequency of the ME composite. The desired unbiased frequency depends largely on the biasing scheme. For example, if the external bias is bipolar, then it may be desirable for the unbiased FMR frequency to be centered on the total operating bandwidth. However, if a monopolar bias (eg. 0 to 5V or 0 to -5V) is used then it may be desirable for the unbiased FMR frequency to be positioned at the upper or lower end of total operating bandwidth.
6.1.2. **Electrostatically tunable bandstop filter**

An electrostatically tunable band-reject filter device was made with a PZT/YIG/GGG trilayer composite loaded in a microstrip fixture with a 1-mm air gap as depicted in Figure 6.1-1. The total FMR frequency shift within this electric bias range was measured and plotted in Figure 6.1-2, as a function of the magnetic bias field. The FMR frequency shift of the YIG/PZT bilayer shows nonlinear bias magnetic field dependence, which is nearly zero when the bias field is between ±50 Oe, indicating that there is no electrostatically induced FMR shift occurring. This observation is consistent with the low frequency ME voltage coefficients and the hysteresis loop, in which the inverse magnetoelastic effect causes domain activities in the YIG layer. The FMR frequency shifts rise nearly exponentially to ~15 MHz when the YIG material is saturated. After saturation, the tunable bandwidth exhibits a nearly linear increase from 15 MHz at a bias field of 100 Oe to 30 MHz at 1200 Oe and dropped suddenly at a field of ~1300 Oe to about 20 MHz. This nonlinear bias field dependence is a direct consequence of interference between the uniform mode ferromagnetic resonance and other magnetostatic spin waves. The position of the spin wave modes may push the FMR frequency higher or lower, leading to the nonlinear bias field dependence of the FMR frequency shift. Double absorption peaks are observed in the $S_{21}$, indicating the close overlapping of the uniform mode and another magnetic spin wave mode as shown in Figure 6.1-3.
Figure 6.1-1. Depiction of electrostatically tuned microwave band-reject filter with embedded GGG/YIG layer facing signal line.

Figure 6.1-2. Total electrostatically induced FMR frequency shift (in megahertz) versus applied dc magnetic field (in oersted).
Peak tunability was achieved at a magnetostatic bias of $H_{DC} = 880$ Oe. The transmission coefficient ($S_{21}$) versus frequency is shown in, with the electrostatic field at +7.2 kV/cm, 0 kV/cm, and -7.2 kV/cm. This is equivalent to a voltage sweep between +350V to -350V. The device has a peak attenuation of $\sim 60$ dB and a 40-dB rejection band of $\sim 13$ MHz. The maximum insertion loss is $\sim 5$ dB, and the electrostatically tunable range (total FMR frequency shift) is $\sim 30$ MHz, which can be further enhanced by optimizing the properties of the composite, such as the volume ratio of the magnetic phase and the piezoelectric phase [6.6]. This device is relatively simple to construct and enables one to characterize the filtering performance of
various layered composites. The ultimate goal of such a device is to optimize the ME composite to achieve maximum selectivity, attenuation, tunability, and microwave frequency operation, while minimizing the external bias (e.g., zero magnetic (or current) bias, and a low electrostatic bias range (voltage bias < ±10V)).

A possible strategy for developing ME based MMIC filters would be to first construct a voltage driven electrostatically tuned ME filter comparable in performance to a current driven magnetostatically tuned filter. This would demonstrate that the ME device performs just as well or better than existing technology, making it a commercially viable product. The next stage in development would be to gradually reduce the scale of the device and adapting synthesis methods compatible to MMIC based microfabrication. The final goal being to realize voltage driven tunable MMIC filters on a semiconductor chip or multilayer MMIC from ME composites.

One of the limiting factors in a MMIC based construction is the small scale motion of the ME composite when it is stressed. One solution would be to form the ME multilayer on a mechanically elastic membrane. The microfabrication techniques for elastic membrane formation can be adapted from existing MEMs (Micro-electromechanical) device fabrication [6.7]. The bigger challenge however is synthesizing piezoelectric layers using a compatible low temperature method. This is the hope of utilizing nanofabrication methodologies [6.1].
6.2. Metamaterial absorbers and filters

An effective electromagnetic wave absorber must fulfill the following requirements: (1) a large part of the incident wave is absorbed while minimizing reflection (2) the absorber must dissipate the incident wave energy (usually by converting it into heat energy). Additional requirements for radar are that the absorber must be thin, flexible, lightweight, and durable. High broad band absorption with low reflectivity (< -20 dB) is generally desired for most applications as in Figure 6.2-1. In the EMI shielding industry reflectivity is sometimes referred to as shielding effectiveness (SE), where $SE = |\text{reflectivity}|$, such that SE > 20 dB is desirable. Additional requirements for EMI shielding applications include high absorption over angle of incidence, high absorption with EM wave polarization (eg. perpendicular, parallel, circular), and high resistivity (for sufficient control of eddy current loss). Wave absorbers however are usually optimized to satisfy a subset of the requirements, at the expense of others for reasons of cost, weight, and size. Figure 6.2-2, includes a table for a pyramidal foam rf absorber. The table shows that the thickness and weight of the material increases with absorption and absorbing bandwidth (expressed in terms of reflectivity). This is a common trait of wave absorbers. Energy absorbed into the material must be sufficiently dissipated to prevent transmission through the material or re-radiation (backscattering) back to the radiating source; the thicker the absorber the more energy it can dissipate. Since energy is normally dissipated as heat, the absorber is also rated by its thermal capacity (or power handling capability). For applications where absorption is critical, thicker
absorbers are used (at the expense of size and weight), while for applications where thickness is important, thinner materials are used at the expense of absorptivity level. However, for applications somewhere in-between, high absorption efficient materials are needed which can be costly to produce. Table 6.2-1 and Table 6.2-2 provide a summary of previous research in wave absorbing materials. EM wave absorbers are generally made up of metallic particles that are used to scatter the wave energy throughout the absorber, a high resistivity material that efficiently converts the energy into heat, and a medium that efficiently dissipates the heat. For applications requiring absorption at a specific frequency, magnetic particles are introduced and optimized to produce strong EM wave absorption at a specific resonant frequency. Another method for wave absorption involves the use of magnetic multilayer structures consisting of impedance matching layers to minimize reflections and absorbing layers to maximize absorption in thin substrates. It is common to introduce a reflecting layer not only to ensure no transmission, but also to maximize the magnetic component of the incident field, so that the adjacent magnetic absorbing layers absorb most of the wave energy (Figure 6.2-3).
Figure 6.2-1. Reflectivity (Return Loss) of Ppy fabric wave absorber [6.8].

![Graph showing reflectivity of Ppy fabric wave absorber]

**Table 1**

<table>
<thead>
<tr>
<th>GRADE</th>
<th>HEIGHT (in. mm)</th>
<th>WEIGHT (lbs, kg)</th>
<th>TIPS PER PIECE</th>
<th>REFLECTIVITY AT FREQUENCY (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SFC-1</td>
<td>3.75 (95)</td>
<td>2.75 (1.25)</td>
<td>255</td>
<td>0.040 0.10 0.3 1.0 3.0 6.0 18 36 56</td>
</tr>
<tr>
<td>SFC-4</td>
<td>4.3 (110)</td>
<td>3.1 (1.4)</td>
<td>144</td>
<td>0.30 0.35 0.42 0.50 0.50 0.50 0.50</td>
</tr>
<tr>
<td>SFC-6</td>
<td>6 (152)</td>
<td>3.5 (1.6)</td>
<td>100</td>
<td>0.32 0.40 0.45 0.50 0.50 0.50 0.50</td>
</tr>
<tr>
<td>SFC-8</td>
<td>8 (203)</td>
<td>4.5 (2.0)</td>
<td>54</td>
<td>0.39 0.47 0.45 0.50 0.50 0.50 0.50</td>
</tr>
<tr>
<td>SFC-12</td>
<td>12 (305)</td>
<td>6 (2.7)</td>
<td>36</td>
<td>0.35 0.40 0.45 0.50 0.50 0.50 0.50</td>
</tr>
<tr>
<td>SFC-16</td>
<td>16 (407)</td>
<td>12 (5.4)</td>
<td>16</td>
<td>0.30 0.37 0.40 0.45 0.50 0.50 0.50 0.45</td>
</tr>
<tr>
<td>SFC-24</td>
<td>24 (610)</td>
<td>17 (7.7)</td>
<td>9</td>
<td>0.30 0.34 0.40 0.45 0.50 0.50 0.50 0.45</td>
</tr>
<tr>
<td>SFC-36</td>
<td>36 (911)</td>
<td>24 (10.0)</td>
<td>4</td>
<td>0.39 0.47 0.42 0.37 0.20 0.10 0.50 0.50 0.50</td>
</tr>
<tr>
<td>SFC-40</td>
<td>40 (1016)</td>
<td>29 (13.5)</td>
<td>4</td>
<td>0.36 0.40 0.37 0.29 0.16 0.07 0.50 0.50 0.50</td>
</tr>
<tr>
<td>SFC-48</td>
<td>48 (1219)</td>
<td>38 (17)</td>
<td>2</td>
<td>0.33 0.28 0.20 0.13 0.05 0.03 0.45</td>
</tr>
<tr>
<td>SFC-60</td>
<td>60 (1524)</td>
<td>43 (19.0)</td>
<td>1</td>
<td>0.20 0.24 0.18 0.12 0.05 0.03 0.45</td>
</tr>
<tr>
<td>SFC-72</td>
<td>72 (1829)</td>
<td>45 (20.5)</td>
<td>1</td>
<td>0.27 0.28 0.20 0.13 0.05 0.03 0.45</td>
</tr>
<tr>
<td>SFC-12</td>
<td>12 (305)</td>
<td>6 (2.7)</td>
<td>36</td>
<td>0.35 0.40 0.45 0.50 0.50 0.50 0.50 0.45</td>
</tr>
<tr>
<td>SFC-16</td>
<td>16 (407)</td>
<td>12 (5.4)</td>
<td>16</td>
<td>0.30 0.37 0.40 0.45 0.50 0.50 0.50 0.45</td>
</tr>
</tbody>
</table>

Figure 6.2-2. Broadband pyramidal rf wave absorber [Cumming Microwave].
Table 6.2-1. Overview of wave absorbing composites. [6.9-6-16]

<table>
<thead>
<tr>
<th>Composite</th>
<th>Primary Benefit</th>
<th>Primary Limitation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bulk Ferrite</td>
<td>Strong wave absorbing properties ($\mu''$)</td>
<td>Snoek's Limit</td>
</tr>
<tr>
<td>Ferrite-polymer</td>
<td>More flexible than bulk ferrite/ceramics</td>
<td>Snoek's Limit</td>
</tr>
<tr>
<td>Ferrite-Metallic Fiber-polymer</td>
<td>Tunability of peak absorption frequency ($\kappa''$)</td>
<td>Narrow bandwidth</td>
</tr>
<tr>
<td>Metal Magnetic Thin Film</td>
<td>Higher Snoek's Limit than ferrites</td>
<td>Eddy current loss</td>
</tr>
<tr>
<td>Metal Magnetic Powder-polymer</td>
<td>Higher Resistivity (addresses eddy current loss)</td>
<td>Reduced Permeability (due to nonmagnetic content)</td>
</tr>
<tr>
<td>Metal Magnetic Powder-Carbon NanoFiber-polymer</td>
<td>Better tunability of complex permeability and permittivity</td>
<td>Poor absorption</td>
</tr>
<tr>
<td>Metal Magnetic Powder-Ferrite Powder-polymer</td>
<td>Improved microwave absorption over metallic magnetic powder-polymer</td>
<td>Better absorption</td>
</tr>
<tr>
<td>Metal Magnetic Powder encapsulated in Carbon Nanotubes</td>
<td>Improved broadband microwave absorption (better match between $\mu''$ and $\kappa''$)</td>
<td>Complex fabrication</td>
</tr>
<tr>
<td>Metamaterials</td>
<td>Composed mainly of metallic structures (simpler fabrication)</td>
<td>Narrow bandwidth</td>
</tr>
</tbody>
</table>
Table 6.2-2. -20 dB reflectivity bandwidth and peak reflectivity of thin film composite wave absorbers. [6.9-6-16]

<table>
<thead>
<tr>
<th>Composite</th>
<th>Film Thickness (mm)</th>
<th>Bandwidth (GHz)</th>
<th>Min Loss at Peak Frequency (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>CINF/MMPFBHSS</td>
<td>5.23</td>
<td>13 to 16</td>
<td>-26</td>
</tr>
<tr>
<td>LiZn ferrite-polymer</td>
<td>5</td>
<td>0.3 to 1.5</td>
<td>-50</td>
</tr>
<tr>
<td>short metal fiber incorporated into ferrite/resin</td>
<td>4.6</td>
<td>8 to 13</td>
<td>not available</td>
</tr>
<tr>
<td>carbon fiber and nanotube based composites with polypyrrole fabric</td>
<td>8.2</td>
<td>8 to 18</td>
<td>-33</td>
</tr>
<tr>
<td>α-Fe/Z-type Ba-ferrite nanocomposite</td>
<td>3.3</td>
<td>8 to 11</td>
<td>-28</td>
</tr>
<tr>
<td>Ni/hexagonal ferrite/polymer composites:</td>
<td>2</td>
<td>8 to 12</td>
<td>-70</td>
</tr>
<tr>
<td>carbon-coated nickel nanocapsules</td>
<td>2</td>
<td>12.5 to 13.5</td>
<td>-32</td>
</tr>
</tbody>
</table>

- **Basic Multilayer absorber:**
  - 1<sup>st</sup> Layer: strong magnetic loss materials
  - 2<sup>nd</sup> Layer:
    - strong electric loss [larger thickness]
    - both electric and magnetic loss materials [smaller thickness]
  - 3<sup>rd</sup> Layer: Impedance Matching Layer or Impedance Transformer

Figure 6.2-3. Absorption through a multilayer absorber. Layer 1 commonly deposited on a reflector (or conducting layer) so that EM wave is either reflected or absorbed.
A relatively new area in wave absorption research is in the utilization of metamaterials. This section will examine the use of metamaterial multilayers for narrowband EM wave absorption, based on the pairing of electric \((ER)\) and magnetic \((MR)\) resonators to independently tune the constitutive parameters \((\mu_{MR} and \epsilon_{ER})\) of the multilayer structure. Where \(\mu_{MR}\) and \(\epsilon_{ER}\) refer to the permeability of the magnetic resonator and permittivity of the electric resonator respectively. By decoupling the electric and magnetic components the expected result is that \(\mu_{ER} = \mu_o\) and \(\epsilon_{MR} = \epsilon_o\). The most common method of designing metamaterials is to use a numerical approach to solve Maxwell’s equations for a given metamaterial pattern, and then numerically extract the effective constitutive parameters [6.18]. Well known effective medium approaches such as the Drude-Lorentz model [6.19] can be used to form an initial metamaterial design from which one can intuitively develop a workable solution, but not predict the ultimate frequency dependent form that the actual parameters take (primarily due to the complex effects of spatial dispersion within the pseudo-homogenized metamaterial) [6.18]. The initial design in this study is based on an L-C-L electric resonator structure originated by Rosen in 1988 [6.20] for use in frequency selective surfaces, then later adapted by Smith et. al [6.21], to design a metamaterial electric resonator. The resonant frequency which can be expressed as

\[
f_o = \frac{1}{2\pi} \sqrt{\frac{2}{LC}}
\]  

(where the 2 in the numerator is due to the two inductive elements in parallel), and thus a relatively large \(LC\) product is required to decrease the resonant frequency to below 10 GHz. The electric coupling strength is dictated by a single capacitive element. Attempts to increase the capacitance by narrowing the gap or
widening the capacitor have been shown to have the detrimental effect of reducing the
coupling strength [6.21] so that much of the incident wave is transmitted.
Alternatively, the capacitive element could be optimized for maximum electric
coupling, while the inductive elements could be tuned to lower the resonant frequency.
However, a single stripe conducting element does not produce significant inductance,
and two inductors in parallel reduces the effective inductance by approximately one
half \( L_{\text{eff}} = \frac{1}{1/L + 1/L} = \frac{L}{2} \). Dr Smith [6.21] proposed the use of rectangular spirals to
address this issue. This report will examine the feasibility of using spiral inductors in
the L-C-L electric resonator arrangement.

For comparison, an EM wave absorber based on Padilla’s design (Figure 6.2-4)
was simulated. The simulated results (Figure 6.2-5), show strong absorptivity of 84
% at ~9.6 GHz, which differs from the published result of 88 % at 11.5 GHz. This
discrepancy is believed to be due slight differences in the properties used to define the
FR4 substrate. The publication did not provide details on the specific constitutive
properties of the FR4 material used, and the dielectric constant of FR4 can vary
around the value of 4. The simulation presented uses the default dielectric constant
value provided in HFSS. As described in section 1.5, the structure consists of two
gapped capacitors in parallel with a stripe inductor. The electric component of the
EM wave couples strongly to the two capacitors at resonant frequency for which the
transmissivity is minimal \( T \to 0 \). This would normally result in high reflectivity
\( R \to 1 \) due to re-radiation as in Figure 6.2-6, however the addition of a cut wire
stripe at the bottom of the FR4 substrate, is used (in conjunction with the center stripe
inductor in the electric resonator structure on the top) to match the resonant impedance to the incident wave impedance

$$Z_{\text{waveabsorber}} = \sqrt{\frac{\mu_{\text{MR}}}{\varepsilon_{\text{ER}}}} = Z_0 \sqrt{\frac{\mu_r}{\varepsilon_r}} = Z_0 = Z_{\text{incident}}$$

to minimize reflectivity $R \to 0$, and resulting in a high absorptivity (see Figure 1.2-1) $A = 1 - R - T \approx 1$. The parallel stripes sandwiching the FR4 substrate couple to the magnetic component of the EM wave resulting in anti-parallel currents. The length, width, and spacing of these two inductive elements determine the magnetic resonance frequency ($f_{\text{resM}} = \frac{1}{2\pi} \sqrt{\frac{1}{L_{\text{MR}}C_{\text{MR}}}}$) and coupling strength. The capacitor width, and gap determine the electric resonance frequency ($f_{\text{resE}} = \frac{1}{2\pi} \sqrt{\frac{1}{L_{\text{ER}}C_{\text{ER}}}}$) and coupling strength. The two capacitors combine ($C_{\text{eff}} = C + C = 2C$) to lower the resonance frequency $f_{\text{resE}}$, but the width of each capacitor is kept small enough to maintain strong coupling. Padilla et. al. [6.22] has performed studies on electric ring resonators with one and two pairs of capacitors in an attempt to optimize the coupling strength per unit area of the unit cell. The magnetic resonator must then be appropriately tuned to match the resonance of the electric resonator so that $f_{\text{resM}} = f_{\text{resE}}$, which implies that the relative permeability and permittivity of the metamaterial wave absorber be equal at the incident plane ($\sqrt{\frac{\mu_{\text{mr}}}{\varepsilon_{\text{mr}}}} = 1 \Rightarrow \mu_{\text{mr}} = \varepsilon_{\text{mr}}$).
Figure 6.2-4. Metamaterial EM plane wave absorber unit cell based on Padilla’s design. FR4: $\varepsilon_r=4.4$, loss tangent=0.02.

Figure 6.2-5. Percent absorption of metamaterial narrowband wave absorber (in figure above).
Figure 6.2-6. Metamaterial bandstop filter: incident wave is re-radiated back to the source at ~6 GHz. (red) return loss [dB] (blue) insertion loss [dB].

Spiral inductors are the most commonly used rf inductors in MMIC designs due to the relative ease of integration, and the extensive amount of effort put into design, fabrication, and modeling of these structures. The inductance can be modified by adjusting the number of turns, and conductor cross section (as in Figure 6.2-7).
Some of the major drawbacks of spiral inductors are turn dependent self-resonant frequency, feed-thru parasitic loss, and mutual inductance loss [6.23, 6.24]. In order to generate more inductance the number of turns is usually increased per the relation: \( L = \frac{N\Phi}{I} \), where \( N \) is the number of turns, \( \Phi \) is the magnetic flux, and \( I \) is the current through the inductor. As the number of turns increase the amount of semiconductor real estate increases, which adds to the chip cost. In addition, the self-resonant frequency (SRF) defined as: \( SRF = \frac{1}{2\pi\sqrt{LC}} \) [6.25] decreases, which limits the operable range of the inductor (Figure 6.2-8). In order to connect the center terminal of the spiral to the rest of the circuit, a cross-over (or cross-under) conductive trace is formed. The cross-over metal forms a bridge over the spiral turns, introducing unwanted parasitic capacitive, resistive, and inductive coupling [6.24, 6.26]. This reduces the quality factor (Q-factor) of the inductor. The Q-factor is a measure of energy storage relative to energy loss: \( Q = \frac{X}{R} \). The higher the Q-factor the more energy efficient the inductor is. Another loss mechanism that impacts the Q-factor is mutual inductance loss. Mutual inductive coupling occurs when the magnetic flux generated from one turn couples with flux generated from an adjacent turn. The flux lines counteract with each other effectively reducing the overall inductance that could have been achieved without the coupling effect.
Figure 6.2-7. Spiral inductors tested at RF/Microwave frequencies to evaluate use in metamaterial resonator. Results shown in Figure 6.2.8.
Figure 6.2-8. Broad band spectrum of spiral inductors. Incorporation of spiral inductors in metamaterial resonator dramatically increases the effective inductance at the expense of self-resonance.

A Rogers 4350 30 mil thick substrate was used which has a loss tangent of 0.004 and a dielectric constant of 3.66. A simulation of the substrate alone in Figure 6.2-9 shows that most of the incident wave is transmitted through the substrate. An L-C-L electric resonator (Figure 6.2-10) was designed with a unit cell size $p$ of 5 mm so
that the free space homogeneity condition \( p < \frac{\lambda}{4} = \frac{30\text{mm}}{4} = 7.5\text{mm} \) is satisfied at the maximum operating frequency of 10 GHz. The electric resonator structure consists of a gapped capacitor with a thickness of 40 µm, which is four times the thickness of the conducting lines, in order to enhance the electric coupling strength (Figure 6.2-11). The inductive elements consist of a pair of 2-turn spiral inductors with feed-thru lines approximately 20 µm above the spiral turns. The simulated results are shown in Figure 6.2-12. Maximum coupling occurs at \( \sim 6.4 \) GHz, and a small amount of absorption is observed which is approximately 18 %. The frequency response is asymmetric, showing a roll off above 6 GHz, resulting in an asymmetric reflectivity and transmissivity spectrum. Examination of the electric coupling strength at resonance (Figure 6.2-13) shows that the electric field component of the EM wave also couples to the feed through parasitic capacitance of the spiral inductors. This is believed to be the source of the asymmetry in conjunction with the SRF.
Figure 6.2-9. EM Wave Absorptivity Characteristics of Rogers 4350 substrate (only). RO4350: $\varepsilon_r=3.66$, loss tangent=0.004.
Figure 6.2-10. Electrically coupled LC resonator unit cell consisting of a pair of two-turn spiral inductors and a rounded gapped capacitor. The conductor width ($W_{ELC}$) is ~0.085 mm, and $p=5$ mm, $s_{coil}~0.128$ mm, $r_{coil}~0.6$ mm.
Figure 6.2-11. Close up of 2-turn spiral inductor crossover and a rounded gapped capacitor. Capacitor width is ~0.44 mm, gap is 20 µm, and height is 40 µm.

Figure 6.2-12. Absorptivity of electrically coupled LC resonator metamaterial consisting of a pair of two-turn spiral inductors and a gapped capacitor.
The electric resonator was then redesigned (Figure 6.2-14 and Figure 6.2-15) to down shift the resonant frequency, by replacing the gapped capacitor with a metal-insulator-metal (MIM) capacitor to boost the absorptivity and electric coupling strength. The MIM capacitor has a surface area of $\pi r^2 = \pi (0.222\, mm)^2 \approx 0.155\, mm^2$ and a gap spacing of 20 $\mu$m, so a stronger electric coupling field was expected. The simulation results of the MIM based metamaterial (Figure 6.2-16) show that the
resonance has down shifted to ~5.3 GHz, relative to the gapped capacitor based metamaterial, while the electric coupling strength noticeably increased (Figure 6.2-17). A check of the magnetic coupling strength (Figure 6.2-18) confirms that the electric resonator unit cells have minimal response to the magnetic component of the incident EM wave. A cut wire structure (Figure 6.2-19) was then added to the bottom of the substrate to form the magnetic resonator as shown in Figure 6.2-20. The length and width is ~4.8 mm and 0.34 mm (which is 4 times the Cu width of the electric LC resonator) respectively. The magnetic coupling behavior is more complex than in the Padilla structure in that surface currents are induced on both spiral elements plus the cut wire as the magnetic wave component passes laterally through the substrate. The induced surface currents on the inductors are parallel to each other, while the cut wire surface currents on the cut wire are anti-parallel to the inductor currents current. With the addition of the cut wire, the wave absorber has absorptivity ~40 % as shown in Figure 6.2-21. It is also apparent in Figure 6.2-22 that the electric coupling in the metamaterial has increased in strength.
Figure 6.2-14. Electrically coupled LC resonator unit cell consisting of a pair of two-turn spiral inductors and a metal-insulator-metal capacitor.
Figure 6.2-15. Close up of 2-turn spiral inductor crossover and a metal-insulator-metal capacitor. Capacitor plates are \(~0.44\) mm in diameter, and plate spacing is \(20\) \(\mu\)m.

Figure 6.2-16. Absorptivity of electrically coupled LC resonator metamaterial consisting of a pair of two-turn spiral inductors and a metal-insulator-metal capacitor.
Figure 6.2-17. Magnitude of E-Field at resonance of electrically coupled LC resonator metamaterial consisting of a pair of two-turn spiral inductors and a metal-insulator-metal capacitor. Electric component of EM wave electrically couples to MIM capacitor.
Figure 6.2-18. Magnitude of H-Field at resonance of electrically coupled LC resonator metamaterial consisting of a pair of two-turn spiral inductors and a metal-insulator-metal capacitor. Minimal magnetic coupling observed.
Figure 6.2-19. Introduction of bottom cut-wire layer to unit cell of electrically coupled LC resonator metamaterial consisting of a pair of two-turn spiral inductors and a metal-insulator-metal capacitor to form an metamaterial multilayer EM wave absorber.
Figure 6.2-20. EM wave absorber unit cell of electrically coupled LC resonator metamaterial consisting of a pair of two-turn spiral inductors and a MIM capacitor to form an metamaterial multilayer EM wave absorber. Addition of cut-wire to increase absorption.

Figure 6.2-21. Absorptivity of metamaterial wave absorber with MIM capacitor.
An equivalent circuit model of the wave absorber was simulated in advanced design system (ADS) as shown in Figure 6.2-23. The model reveals that the roll-off characteristic observed beyond 6 GHz is linked to losses in the parallel inductors. The asymmetric spectrum is thus believed to be due to the effect of the feed-through capacitance along with other parasitic losses in the spirals. The behavior is enhanced
by the increase in parasitic coupling of the feed through lines of the inductors as observed in Figure 6.2-22.

Figure 6.2-23. Equivalent circuit model reveals that the roll off characteristic beyond 6 GHz may be linked to losses in the parallel inductors. (Simulation performed using ADS).

The resonant frequency was further reduced to ~4.1 GHz by reducing the dielectric spacing of the capacitor from 20 µm to 10 µm. A second cut-wire was added to maintain the absorption (Figure 6.2-24). By tuning the width of the cut-wire to 6.5 times the width of the conducting paths of the electric LC resonator unit cells \( (W_{ELC}) \) the absorption peaked at about 46 % (Figure 6.2-25). Smaller 1-3 % enhancements may be achieved using a lossier substrate (substrate loss) or conductive
material (conductive loss). However previous studies have concluded that dielectric loss is the main absorption mechanism. Therefore, increasing the electric coupling strength is necessary for a more dramatic enhancement. By minimizing the effect of the feed-through capacitance, increasing the SRF, and optimizing the electric coupling strength, this design is expected to produce a more favorable symmetric reflectivity, transmissivity, and absorptivity spectrum, with stronger absorption.

Figure 6.2-24. EM wave absorber unit cell of electrically coupled LC resonator metamaterial consisting of a pair of two-turn spiral inductors and a MIM capacitor to form a metamaterial multilayer EM wave absorber (2-cutwires).
6.2.1. Magnetic loading using microwave magnetic films

Magnetic films can be introduced to increase the magnetic flux density of the spirals, reducing the number of turns required to achieve the desired inductance, and minimize mutual inductance. Losses (dielectric, conductive, etc) from the magnetic films however, must be controlled in order to maximize the Q-factor. In many cases the Q-factor is reduced [6.27]. The change in direction of the rf field within the spiral frustrates the directional anisotropy in the magnetic film further reducing the Q-factor.
A dielectric magnetic film, with no internal directional anisotropy such as a ferrite could be used to minimize some of these key loss mechanisms, however, the high temperature film growth process limits its use on temperature sensitive substrates. A low temperature growth process called spin-spray deposition may be adapted for synthesizing ferrite films [6.28-6.29]. It opens the door to utilizing ferrites in more semiconductor based applications. A spin-spray system shown in Figure 6.2-26, was used to deposit spin-sprayed NZFO ferrite films on unpolished alumina as shown in Figure 6.2-27. Spin spray films were also successfully deposited on a silicon substrate. There however, remain issues with producing high quality self-biased spin-spray films to eliminate the need to apply an external magnetic bias to saturate the film. The relative permeability is also comparably smaller (~35 for NZFO films deposited) than most metallic magnetic films (several hundred). A metamaterial wave absorber is shown conceptually with self-biased 20 µm thick spin-spray ferrite films in Figure 6.2-28.
Figure 6.2-26. Ferrite Spin-Spray System. Low temperature deposition of ferrite films.

Figure 6.2-27. Spin-sprayed spiral inductors on 2mm unpolished Alumina substrate. Conducting spiral layer (10 nm Ti/3 μm Cu) is sandwiched between two layers of 2.5 μm NZFO.
Figure 6.2-28. Application of self-biased spin-spray ferrite films to enhance inductance in metamaterial wave absorber.

Magnetic stripe inductors are planar conducting lines with a stripe of magnetic film deposited above, below (Figure 6.2-29 and Figure 6.2-30), or enclosing the
conductor. Much like a spiral inductor, as the rf signal passes through the conductor, a magnetic flux circulates around the conducting lines and is partially (or fully if magnetic film surrounds the conductor) enhanced by the magnetic multilayer film beneath the copper line. The key advantages of using stripe inductors over spiral inductors is the elimination of the feed-thru line to complete the circuit, and elimination of mutual inductances between turns. The stripe inductor is a single turn inductor therefore a large flux density must be generated in the stripe to compensate for the lack of turns. Thus the inductance is enhanced primarily by the magnetization of the magnetic film requiring a multilayer structure like shown in Figure 6.2-31. Multilayer FM/AFM/FM metallic magnetic films are an attractive choice for its high magnetization property, and enhanced operating range due to extended FMR frequency. To achieve high inductance/length the magnetic multilayers must be made thick enough to meet the inductance requirements, while minimizing conductive losses related to the skin effect $\delta = \sqrt{\frac{\pi f \sigma}{\mu}}$. A single unit cell of a metamaterial EM wave absorber, consisting of metallic magnetic multilayer stripe inductors, is depicted in Figure 6.2-32.
Figure 6.2-29. 2-port stripe inductor: 4 µm multilayer film/3 µm copper film.

Figure 6.2-30. Transmission lines with 8-period magnetic multilayer stripes. 2 mm Alumina/1 nm Ti/4 µm FeCoB multilayer/3 µm Copper. Magnetized striped inductive elements may be incorporated into L-C-L electric resonator to reduce resonance frequency of unit cell when operated below SRF of magnetic multilayer.
Figure 6.2-31. 8-period FeCoB metallic magnetic multilayer structure.

Figure 6.2-32. Unit cell of metamaterial wave absorber, consisting of electric LC resonator with metallic magnetic multilayer stripe inductors. Magnetic stripes eliminate the effects of feed through capacitance, while reducing resonant frequency when metamaterial is utilized below the FMR of the multilayer film.
In conclusion, metamaterial EM plane wave absorbers are ideally suited for microwave applications requiring selective narrowband energy absorption. They can be designed to operate at any frequency by independent tuning of the electric and magnetic resonating elements, and can be optimized for a wide range of dielectric substrates and conducting materials. A potential area of development would be in its use in EM coupling, filtering, or noise suppression microwave systems. The first step in this direction would be to develop metamaterial wave absorbers with unit cell and lattice size comparably smaller than the elements in the microwave system, in addition to the operating wavelength. For applications below 5 GHz, metallic multilayer and spin-spray magnetic films may be incorporated to boost the effective inductance of metamaterial LC resonator unit cells, lowering the resonant frequency. Alternatively, the enhanced inductance/area allows the metamaterial elements to be scaled down in size. Choosing MIM capacitors over edge-gapped capacitors significantly boosts the LC capacitance and quality factor. Introducing dielectric materials between the MIM plates allows for further enhancement of capacitance/area of the metamaterial wave absorber.

6.3. References

Chapter 7. Summary

In summary, metallic magnetic layered composites are well suited for the fabrication of MMIC scale devices due to the ability to produce high quality rf sputtered nanometer scale thin films with self-bias properties ($H_r < 2$ Oe) that eliminate the need for an external magnetic bias. High magnetization and narrow linewidth capability allow for the incorporation of high inductances per unit surface area by precisely controlling layer thickness, composition, magnetic anisotropy, and periodicity of layered microwave magnetic composites. Metallic magnetic multilayer films are cost effective in that they can be readily adapted to existing planar microfabrication processes. RF sputtering allows for low temperature deposition on many microwave substrates such as silicon, alumina, and GaAs (as well as many other III-V) materials by introducing suitable seed layers (e.g. Ti, Fe, Ta) and insulating layers (AlO$_3$) for the target application. In addition to substrates, the films can be deposited on existing layers or structures, making them compatible to bottom-up MMIC fabrication.

Ru-seeded CoFe/PtMn/CoFe (ferromagnetic/antiferromagnetic/ferromagnetic) trilayer composites show excellent magnetic softness with a low hard axis coercivity of 2–4 Oe, an easy axis squareness ratio ($M_r / M_s$) of $>98 \%$, a significantly enhanced in-plane anisotropy of 123 Oe, and a low ferromagnetic resonance (FMR) linewidth of 45 Oe at $\sim$9.5 GHz when the CoFe layer thickness is above 200 Å. With the combination of these magnetic and microwave properties, the CoFe/PtMn/CoFe films
could play a valuable role in microwave applications by extending the operating frequency of metallic magnetic composite based devices such as rf inductors and filters. These films could potentially be used as the magnetic phase in magneto-electric (ME) layered composites to optimize the unbiased center frequency for a given application.

A set of broad band characterization techniques and apparatus were developed specifically for ME bilayer characterization. A microstrip technique involving the application of an electrostatic bias sweep at different magnetostatic fields was used to evaluate the broad band performance of a YIG/PZT bilayer that reveals non-linear characteristics at low and high magnetostatic fields. The negligibly small FMR frequency shift at low frequencies was due to the magnetic domain activities; meanwhile, the magnetostatic spin waves were responsible for the nonlinear bias field dependence of the FMR frequencies at high magnetic bias fields. The same technique for broadband testing of ME bilayer composites was also used to fabricate a magnetostatically tunable band-reject filter. This characterization technique can be extended to other types of layered ME composites as a means of studying the electrostatic tunability as a function of magnetostatic bias to produce a comparative figure of merit between materials. A CPW technique involving the use of a PZT beam structure was developed for characterizing the change in permeability of FeCoB/PZT and FeGaB/PZT composites. This is particularly useful in applications such as tunable inductors and phase shifters, as well as filters. Strong ME coupling at microwave frequencies have been demonstrated in these metallic magnetic film/PZT
ME composites. Large FMR frequency shifts of 50 and 110 MHz are observed in the FeCoB/PZT and FeGaB/PZT composites, respectively, when tuned with an electrostatic field of ±8 kV/cm. Large FMR frequency tunability of 28% at 2.5 GHz is predicted for FeGaB/PZT ME composite when FeGaB is deposited directly on PZT, which provides tunable microwave ME devices with large tunability.

Metamaterial wave absorbers were designed and simulated in HFSS. It involves utilizing a multilayer structure that isolates the electric coupling from the magnetic coupling to absorb radiation from an incident electromagnetic plane wave. Self-oscillating electric resonators utilizing spiral inductors and metal-insulator-metal capacitors were examined with the intent of maximizing unit cell absorption in order to minimize cell size (and lower the resonant frequency of the resonator). When paired with a magnetic resonator with matching resonance, a polarization dependent narrowband plane wave absorber was designed. The absorber achieved a maximum absorptivity of 46%, at ~4GHz.