IMPLANTABLE FERRITE ANTENNA
FOR
BIOMEDICAL APPLICATIONS

by

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ABSTRACT

We have developed an implantable microstrip patch antenna with dimensions of $10 \times 10 \times 1.28$ mm, operating around the Industrial, Scientific and Medical (ISM) band (2.4 – 2.5 GHz). The antenna is characterized in skin-mimicking gels and compared with simulation results. The experimental measurements are in good agreement with simulations, having a -16 dB reflection coefficient and -18 dBi realized gain at resonance, with a 185 MHz -10 dB bandwidth. The simulated effects of ferrite film loading on antenna performance are investigated, with comparisons made for 5 and 10 μm thick films, as well as for 10 μm thick films with varying magnetic loss ($\tan \delta_\mu = 0.05, 0.1$ and 0.3). Our simulations reveal that the addition of 10 μm thick magnetic layers has effectively lowered the resonant frequency by 70 MHz, while improving return loss and -10 dB bandwidth by 3 dB and 40 MHz, respectively, over the uncoated antenna. Ferrite film coating also improved realized gain within the ISM band, with largest gain increases at resonance found for films having lower magnetic loss. Additionally, the gain (G) variance at ISM band limits, $\Delta G_{f(2.5\text{GHz})-f(2.4\text{GHz})}$, decreased from 1.97 to 0.44 dBi for the antenna with 10 μm films over the non-ferrite antenna.

The measured dip-coated NiCo ferrite films effectively reduces the antenna resonance by 110 MHz, with a 4.2 dB reflection coefficient improvement as compared to an antenna without ferrite. The measured ferrite antenna also reveals a 6 dBi and 35 MHz improvement in realized gain and -10 dB bandwidth, respectively, at resonance. Additionally, the ferrite-coated antenna shows improved directivity, with wave propagation attenuated at the direction facing the body

internal. These results indicate that implantable antenna miniaturization and reliable wireless communication in the operating frequency band can be realized with ferrite loading.

DEDICATION

This thesis is dedicated to all of those who have brought me guidance, comfort and laughter throughout this endeavor. I express my sincerest gratitude to my family and friends who support and inspire me on a daily basis. My comrades in the Magnetic Materials and Device Laboratory have been true allies throughout my graduate career and I could not have completed this work without them. Finally, I would like extend my appreciation to my advisor, Dr. Yang-Ki Hong, for the consistent support and direction he has eagerly provided throughout my graduate studies.

LIST OF ABBREVIATIONS AND SYMBOLS

2-MOE	2-methoxy-ethanol
θ	Angle of diffraction
a.u.	Arbitrary units
ϕ	Azimuthal angle
f_c	Center frequency
Δ	Change of quantity
CVD	Chemical vapor deposition
Co	Cobalt (element)
σ	Conductivity (electrical)
d	d-spacing
dB	Decibel
dB _i	Decibel isotropic
°C	Degrees Celsius
$\tan \delta_\epsilon$	Dielectric loss tangent
D.I.	Deionized (water)
DC	Direct current
D	Directivity
ECLA	Electrically coupled loop antenna
emu	Electromagnetic units
θ	Elevation angle

FCC	Federal communications commission
FMR	Ferromagnetic resonance
FEM	Finite element method
FR4	Flame retardant 4 (printed circuit boards)
G	Gain
GHz	Gigahertz
g	Gram (mass unit)
HFSS	High frequency structural simulator
ISM	Industrial, scientific and medical (radio spectrum)
n	Integer (any)
Fe	Iron (element)
K_{α}	K-alpha radiation
kg	Kilogram (mass unit)
kHz	Kilohertz
$\tan \delta_{\mu}$	Magnetic loss tangent
MHz	Megahertz
MICS	Medical implant communications service
m	Meter (length unit)
μm	Micron (length unit)
ml	Milliliter (liquid volume unit)
mm	Millimeter (length unit)
min	Minute(s) (time unit)
Ni	Nickel (element)

Oe	Oersted (auxiliary magnetic field unit)
Ω	Ohm (electrical resistance unit)
O	Oxygen (element)
PIFA	Planar inverted-F antenna
PEG	Polyethylene glycol
PTFE	Polytetrafluoroethylene
PCB	Printed circuit board
PLD	Pulsed laser deposition
Q	Quality factor
RE	Radiation efficiency
RF	Radio-frequency
RG	Realized gain
Γ	Reflection coefficient
μ_r	Relative permeability
ε_r	Relative permittivity
S	Siemens (electrical conductance unit)
SAR	Specific absorption rate
SMA	SubMiniature version A
3D	Three-dimensional space
VNA	Vector network analyzer
VSM	Vibrating sample magnetometer
γ	Wavelength
XRD	X-ray diffraction

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CHAPTER 1

INTRODUCTION

1.1. Motivation

The field of biomedical devices and systems has experienced a sharp increase in research attention over the last decade. The resulting influx of new and more sophisticated biomedical technologies has inspired a need for efficient methods of communication and power transfer to devices embedded in the human body, such as cardiac pacemakers and neural recording systems. Furthermore, one of the most cumbersome components of implantable devices is the battery unit, replacements of which often require routine, invasive surgical methods. Accordingly, implantable antennas have attracted considerable interest recently for their ability to provide wireless energy to implantable medical devices, with the primary focus of contemporary research concerning antenna miniaturization, efficiency and biocompatibility. Achieving reliable wireless power transfer and data communication is imperative to the efficacy of biomedical implant systems and introduces significant design challenges.

The leading objective regarding wireless communication with implantable devices is to reduce the antenna dimensions while concurrently overcoming degenerative effects resulting from the lossy biological tissues enveloping the system [1]. Various antenna miniaturization techniques have been employed by other authors, including radiator geometry optimizations and high permittivity substrate/superstrate materials [2]. Moreover, shielding the conductive elements of the antenna from contact with human tissues is critical for device efficiency and biocompatibility. Previously reported implantable designs have incorporated biocompatible

housings to shield the antenna from the surrounding tissues [1]. Though many of these techniques are useful, their benefits can be limited by physical constraints. Therefore, alternative or supplemental methods of improving device performance should be explored.

Our proposed method to address these issues is to employ a thin ferrite film coating to the implantable antenna, which can aid in improving antenna isolation and biocompatibility.

Magnetic particles have been widely studied for biomedical applications, including enzyme immunoassay [3], hyperthermia therapy [4], as well as magnetic resonance imaging and drug delivery [5]. Furthermore, ferrite films have garnered recognition for radio-frequency (RF) device miniaturization [6], with specific attention to NiCo iron oxide films [7–8]. In [9], a $\text{Ni}_{0.23}\text{Co}_{0.27}\text{Fe}_{2.50}\text{O}_4$ film is reported exhibiting a relatively high permeability (μ_r) and low magnetic loss ($\tan \delta_\mu$) of 10 and 0.3, respectively, at 2.45 GHz. These magnetic properties, in addition to their high electrical resistivity [10] and low toxicity [11], make ferrite films excellent candidates for application to implantable RF devices.

We have developed a miniaturized microstrip patch antenna operating near the Industrial, Scientific and Medical (ISM) band (2.4 – 2.5 GHz) designed for power transfer and data communication with medical devices implanted in the human body. Our proposed antenna design offers a unique radiator geometry, optimized for efficient power transfer in the desired frequency band of operation. The fabricated implantable antenna is characterized in human tissue-mimicking gels and compared with simulation results. Detailed insight into the antenna design, simulation, optimization, and characterization processes will be given in this work and, most notably, the effects of ferrite film thickness and magnetic properties on antenna performance are investigated.

1.2. Challenges & Objectives

When developing a methodology for implantable antenna design, several parameters must be taken into consideration depending on the application and associated constraints, such as antenna size, return loss, gain, and frequency band of operation. Typically, the *type* of antenna configuration is one of the first aspects to consider. Various classes of implantable antenna configurations have been proposed in the past, each with its own advantages and drawbacks. Two commonly considered antenna types are the helical antenna and electrically coupled loop antenna (ECLA). These configuration types have led to efficient implantable antenna designs by other authors [12–13]; however the non-planar profile of these models can make antenna miniaturization more difficult. With this in mind, we chose the microstrip patch antenna configuration as the preferred starting model for our implantable antenna design.

The microstrip patch antenna is a simple antenna structure composed of a planar substrate, typically rectangular in surface dimensions, with an upper radiating patch electromagnetically coupled to a lower ground plane. This antenna configuration was chosen due to its planar profile, compliance with peripheral circuitry, and the relative simplicity of fabrication [14]. Additionally, the microstrip patch antenna can support a variety of specialized radiator geometries, including square, loop, serpentine, spiral and meandering [2]. Likewise, with the additional of a shorting pin, the microstrip patch antenna can be converted into a planar inverted-F antenna (PIFA) with better performance [2]. An example of a typical *rectangular* microstrip patch antenna is given in Fig. 1.1 [15]. The key parameters to note are the antenna substrate, ground plane, as well as patch radiator and transmission line. Our subsequently proposed implantable antenna will be modeled from this type of antenna configuration, with

optimizations performed to produce a unique, novel radiator geometry to improve antenna miniaturization and performance in the ISM band.

FIG. 1.1. Typical rectangular microstrip patch antenna configuration [15].

The next design parameter to consider is the frequency band in which the device will operate. This is important for a variety of reasons, including the fact that antenna size is inversely proportional to the resonant frequency; typically, the lower the desired frequency of resonance is, the larger the antenna dimensions must be. Additionally, operational frequency bands for wireless devices cannot be chosen just anywhere in the electromagnetic spectrum, because there are specific bands allocated for medical device communication. The two main frequency bands for this application are the Medical Implant Communications Service (MICS) band and the Industrial, Scientific and Medical (ISM) band. These two frequency bands are actually composed of multiple bands that are various ranges of the radio spectrum. The MICS band is primarily composed of MHz frequency bands, including 405 – 406 MHz, 413 – 419 MHz, 451 – 457 MHz, as well as others. The ISM band has several frequency allocations across the majority of radio spectrum, from kHz ranges up to several hundred GHz. For implantable device applications,

there is a tradeoff to consider when determining the appropriate frequency band of operation. The specific absorption rate (SAR), which is a measure of the rate at which the human body absorbs energy when exposed to an electromagnetic field from the RF spectrum, increases with increasing operation frequency. However, as previously mentioned, the antenna dimensions tend to increase with a decrease in required frequency band. Therefore, the tradeoff is between size and SAR, and the ISM band of 2.4 – 2.5 GHz was chosen to reasonably satisfy both parameters. This low GHz band allows for reduced antenna dimensions, and with a low power dissipation will satisfy the SAR regulations imposed by the FCC [2].

Other implantable antenna aspects to consider are antenna dimensions, long term biocompatibility, electrical insulation from the body, signal attenuation from body tissues, as well as overcharge and discharge protection. These issues, as well as others, will be investigated and addressed throughout this work, specifically effects observed by deposition of ferrite films on the antenna substrate.

CHAPTER 2

METHODS & PRINCIPLES

2.1. Antenna Considerations & Simulations

Prior to fabrication and characterization of any antenna models, carefully designed and optimized antenna simulations must first be performed to verify the device is operating as desired. All antenna modeling and simulation was performed with ANSYS High Frequency Structural Simulator (HFSS) software (v. 16). HFSS utilizes the finite element method (FEM) to solve the corresponding Maxwell equations and boundary value problems for the partial differential equations of our system. FEM is a numerical technique that, for our application, will divide the structure into very small, three-dimensional tetrahedral shapes, also known as meshes. These finite elements are represented by manageable equations which are then aggregated to produce a complete system of equations modeling the total system. All of the subsequent antenna simulations are of the discrete, linear-step type, with a frequency step size of 10 MHz.

As an *implantable* antenna, this receiver device is obviously not designed to operate in air, but rather embedded into human tissue at a specific depth. Therefore, in order to model an appropriate simulation environment, the dielectric properties of the biological tissues of interest must be understood. Implantable antenna receivers are most commonly embedded only a few millimeters into a patient's body [1], so as to limit radiation deficiencies caused by the enveloping tissues; that is, a larger implant depth results in lower antenna efficiency due to the dielectric losses associated with the human tissues. Though the human body will be the ultimate

operation location of the antenna, air simulations are also performed in order to facilitate a more accurate antenna validation procedure.

Our implant depth will be approximately a few millimeters, and a typical device implantation location is the area around the human chest, for example for cardiac pacemaker communication. Therefore, the tissues of interest will be human skin (dry), fat and muscle, dielectric properties of which at 2.45 GHz are compiled in Table 2.1 from accepted reports in literature [16–18]. The considered dielectric properties include relative permittivity (ϵ_r), dielectric loss tangent ($\tan \delta_\epsilon$), and electrical conductivity (σ), and the property values in Table 1 represent the simulated values and desired values for the subsequently fabricated tissue-mimicking gels. Relative permittivity is the measure of resistance that is encountered when a material is exposed to an electric field in some non-vacuum medium. Similarly, dielectric loss tangent reveals the innate degree of which a material dissipates electromagnetic energy as heat. In Table 2.1, the values for ϵ_r and $\tan \delta_\epsilon$ of both human skin and muscle are relatively high, explains the difficulty in achieving high efficiency communication with antennas embedded in the human body.

	ϵ_r	$\tan \delta_\epsilon$	σ [S/m]
Skin (Dry)	38	0.2826	1.464
Fat	5.28	0.1452	0.1045
Muscle	52.73	0.2419	1.739

Table 2.1. Dielectric properties for human skin (dry), fat and muscle at 2.45 GHz [16–18].

The electrical conductivity for skin and muscle tissues are also significant, and could cause potential antenna performance and biocompatibility issues if the radiating elements of the device are exposed to these conductive tissues. Therefore, some efforts must be considered in an attempt to insulate the antenna radiator and ground plane from the surrounding tissues. Our intent is to deposit biocompatible ferrite film coating to the conducting elements of the antenna, which will be addressed later in this work. In addition to the dielectric properties of the considered human tissues, the physical dimensions must also be considered to accurately model such tissues.

The average thickness for human skin is approximately 2 mm, and this value was used for the skin portion of our simulation model; there is more complexity, however, in choosing thickness values for fat and muscles these tissues exhibit more variance between different body types. Therefore, we have fat and muscle thickness values for simulations to be 4 mm and 30 mm, respectively. The assumed fat thickness is a good estimate for the thickness of average human subcutaneous fat [19], while the muscle thickness was chosen to be rather large in order to represent a realistic depth of tissue behind the device (towards the internal of the body). Figure 2.1 depicts the complete three-layer tissue model used for simulations, which will later be compared with a single-layer, human skin phantom.

FIG. 2.1. Three-layer tissue phantom (skin/fat/muscle) for HFSS antenna simulations.

The three-layer tissue phantom is a rectangular cuboid with overall dimensions of $100 \times 100 \times 36$ mm, resulting in a total tissue volume of $360,000 \text{ mm}^3$. The boundary conditions for our simulations are imposed by a radiation air-box extending 10 mm beyond each face of the tissue model (i.e. $120 \times 120 \times 56$ mm). A comprehensive list of relevant simulation properties for all modeled materials is given in Table 2.2. To remain consistent with our experimental measurement setup, the implantable antenna designs were also simulated in a 12 mm thick human skin model, shown in Fig. 2.2. This single-tissue model has dimensions of $100 \times 100 \times 12$ mm, with similarly assigned boundary conditions (radiation air-box: $120 \times 120 \times 32$ mm). Details relating to the similarities and discrepancies between the two simulation models will be explained later in this work.

* Properties varied during parametric ferrite simulations

Material	ϵ_r	μ_r	σ [S/m]	$\tan\delta_\epsilon$	$\tan\delta_\mu$	Mass Density [kg/m ³]
Air	1.0006	1.0000004	0	0	0	1.614
Copper	1	0.999991	5.8e6	0	0	8933
Skin (Dry)	38.007	1	1.464	0.28262	0	1020
Fat	5.2801	1	0.10452	0.14524	0	900
Muscle	52.729	1	1.7388	0.24194	0	1060
NiCo Ferrite	12	10*	3.25e-6	0.02	0.3*	3200

Table 2.2. HFSS material properties for simulated air, copper, skin (dry), fat, muscle and NiCo ferrite film.

FIG. 2.2. Human skin model for HFSS antenna simulations (100×100×12 mm).

With the environmental tissue models complete, the antenna structure itself must be simulated. As previously mentioned, the microstrip patch antenna was chosen as our antenna configuration of choice, however, assigning the desired substrate material can be a more

complicated endeavor based on the variety of available options. Dielectric substrate materials with high relative permittivity are often chosen for implantable patch antennas because of their ability to lower the effective wavelength and result in reduced resonance frequencies, thus aiding in antenna miniaturization [2]. For implantable antenna applications, some of the most commonly used commercially available dielectric substrate materials are produced by the Rogers Corporation. Rogers Corporation offers high quality substrate materials with a wide range of dielectric properties to choose from. Additionally, polytetrafluoroethylene (PTFE), the ceramic composite comprising the base material of Rogers' laminates, exhibits a relatively high heat tolerance when compared to other analogous dielectric materials. Next, the preferred model of Rogers material must be considered.

Through preliminary antenna calculations and simulations, we have determined the Rogers material RO3010 to be our desired substrate material, with a standard thickness of 1.28 mm. This laminate has a high dielectric permittivity constant, $\epsilon_r = 10.2$, which will aid in antenna miniaturization and efficiency. This is because the frequency of operation for microstrip patch antennas is inversely proportional to the square root of the substrate material relative permeability, such that the center frequency is approximately given by:

$$f_c = \frac{c}{2L\sqrt{\epsilon_r}}, \quad (2.1)$$

where c is the speed of light (299,792,458 m/s) and L is the length of the patch. However, the increase in miniaturization and efficiency offered by Rogers substrate materials comes with one major drawback: the relatively large costs per sheet, as compared to other commonly used PCB materials, such as FR4. Nevertheless, this increase in production cost is alleviated by the miniaturized dimensions of the antenna, which reduces the volume of Rogers material required for each antenna. The complete set of material properties for the Rogers RO3010 laminate [20] is

listed in Table 2.3, including the parameter testing methods and conditions. Fortunately, ANSYS HFSS software has material properties for a variety of Rogers laminates, including RO3010, pre-programmed into the stock material database.

Property	RO3010	Direction	Unit	Condition	Test Method
Dielectric Constant, ϵ_r , Process	10.2 $\pm$ 0.30	Z	-	10 GHz 23°C	IPC-TM-650 2.5.5.5 Clamped Stripline
⁽²⁾ Dielectric Constant, ϵ_r , Design	11.20	Z	-	8 GHz - 40 GHz	Differential Phase Length Method
Dissipation Factor, $\tan \delta$	0.0022	Z	-	10 GHz 23°C	IPC-TM-650 2.5.5.5
Thermal Coefficient of ϵ_r	-395	Z	ppm/°C	10 GHz -50 to 150°C	IPC-TM-650 2.5.5.5
Dimensional Stability	-0.35 -0.31	X Y	mm/m	COND A	IPC TM-650 2.2.4
Volume Resistivity	10 ⁵		M Ω •cm	COND A	IPC 2.5.17.1
Surface Resistivity	10 ⁵		M Ω	COND A	IPC 2.5.17.1
Tensile Modulus	1500	X, Y	MPa	23°C	ASTM D638
Moisture Absorption	0.05	-	%	D48/50	IPC-TM-650 2.6.2.1
Specific Heat	0.8		J/g/K		Calculated
Thermal Conductivity	0.95	-	W/m/K	50°C	ASTM D5470
Coefficient of Thermal Expansion	13 11 16	X Y Z	ppm/°C	-55 to 288°C	ASTM D3386-94
Td	500		°C TGA		ASTM D3850
Density	2.8		gm/cm ³	23°C	ASTM D792
Copper Peel Strength	9.4		lb/in	1 oz. EDC After Solder Float	IPC-TM-2.4.8
Flammability	V-0				UL 94
Lead Free Process Compatible	YES				

Table 2.3. Material properties for Rogers RO3010 laminate [20].

Before complete antenna simulations and optimizations can be performed, constraints must be placed on our implantable antenna dimensions. Our substrate height, or thickness, has already been determined to be 1.28 mm, based on the standard thickness options for Rogers RO3010 laminates; therefore, only the surface dimensions remain unassigned. Typical volumes

for implantable antennas proposed in the past by other authors range widely depending on the specific application, dielectric substrate material, operating frequency band, as well as implantation location and depth. In fact, A. Kiourti and K. S. Nikita report the volumes for a variety of implantable antennas proposed by various authors, which range from around 100 mm^3 to over $10,000 \text{ mm}^3$; among these, the average volume is on the order of 1000 mm^3 [2]. Though reducing the antenna dimensions is among the top primary design objectives, miniaturization comes at the cost of antenna efficiency. Therefore, we want to limit our antenna dimensions as much as possible, while avoiding serious performance deterioration. With these considerations in mind, and after review of previously reported implantable antenna volumes, we have determined our desired dimensional constraints to be an antenna with a volume between 100 and 200 mm^3 . Since one of the three antenna dimensions is already defined (thickness = 1.28 mm), choosing surface dimensions of $10 \times 10 \text{ mm}^2$ would put our antenna volume (128 mm^3) within our desired range.

With the antenna dimensions, configuration type and substrate material assigned, antenna simulations can be performed to model various radiator geometries in an attempt to optimize antenna efficiency in the desired frequency band of operation. It should be noted that since the ultimate goal of this work is to reduce the antenna frequency, i.e. antenna miniaturization, via ferrite film deposition, the frequency of operation for the *uncoated* antenna should be slightly above the desired band. This is because the hope is that the employment of ferrite coatings will lower this operation frequency to within the ISM band limits. The simulation and optimization methods for our implantable antenna design will be explained later in this work.

2.2. Fabrication Methods

2.2.1. Implantable Antenna Fabrication

The fabrication of our implantable PIFA antenna is relatively straightforward, with only a few significant steps. The patterning of the antenna radiator was performed with a precision printed circuit board (PCB) milling machine (LPKF ProtoMat S62). Likewise, the LPKF was used for drilling of the shorting pin slot as well as the outer antenna edges. Upon final PCB milling, each edge of the antenna was smoothed with a low grit sandpaper. The copper shorting pin of radius = 0.6 mm was then cut to the appropriate length (≈ 1.3 mm), inserted into the slot and soldered to both the antenna radiator and ground plane. For compatibility with measurement equipment, we employed a 50 Ω female SMA coaxial connector by soldering the antenna transmission line and ground plane to the SMA center core and lower ground mounts, respectively. Prior to coaxial connection, the SMA connector was first cut and modified to reduce its volume and accommodate the specific size of this miniaturized antenna.

2.2.2. Development of Skin-Mimicking Gel

To characterize the antenna in conditions resembling human body implantation, skin-mimicking gels were fabricated with procedures modified from [21]. In this process, we attempt to create gels with dimensional and dielectric properties as close to our simulation environment as possible. To begin, 26.5 g of sucrose was added to 24 ml of deionized (D.I.) water, mixing the solution with a magnetic stirrer until the particles completely dissolved. To increase gel viscosity, 0.55 g of agarose was then mixed into the solution until a clear, homogeneous liquid formed. Finally, the mixture was brought to a boil for 4 minutes to activate the agarose. Upon removing the liquid solution from the hotplate, the antenna was implanted into the center of the

phantom, which was then left to cool and solidify at room temperature. The final skin-mimicking gel, shown in Fig. 2.3 is found to be extremely durable, retaining its shape and composition throughout antenna characterization.

FIG. 2.3. Fabricated skin-mimicking gels.

Desired dielectric properties for the skin-mimicking gels at 2.45 GHz ($\epsilon_r = 38$, $\tan \delta_\epsilon = 0.2826$, and $\sigma = 1.464$ S/m) were modeled from [16–18], with relative permittivity and dielectric loss verified using a vector network analyzer (VNA: Agilent N5230A) ($\epsilon_r = 38.45$ and $\tan \delta_\epsilon = 0.39$). The measured dielectric constant of the gel is within error of our desired value (38), however, our measured loss tangent is higher than that of our desired value (0.2826). In the subsequent antenna characterization and analysis processes, this increased dielectric loss will need to be taken into consideration as it will affect the resulting gain of the implantable antenna.

2.2.3. Ferrite Dip-Coating Deposition

To realize the ferrite implantable antenna, ferrite deposition methods should be considered thoroughly, due to the temperature limitations imposed by the substrate material. The melting and decomposition temperatures of polytetrafluoro-ethylene (PTFE) are approximately 327 and 400 °C, respectively [22]; the former representing the maximum allowable temperature for our ferrite film deposition and post-processing. This temperature restriction introduces a significant challenge, as the processing temperatures for typical ferrite film deposition techniques (PLD, CVD, sputtering, etc.) often exceed several hundred °C [23]. Therefore, alternative methods to traditional ferrite film deposition techniques must be considered.

Low-temperature ferrite film deposition procedures are quite limited in number, though electroplating [23] and spin-spray [7–8] methods have been successfully attempted. Both techniques are viable options for implantable antenna ferrite loading, though the sol-gel dip-coating method [24] will be our primary deposition process of interest. Dip-coating is a low-cost deposition technique that can provide fully-encapsulating spinel-type magnetic thin films [25] at relatively low post-processing temperatures. Depositing ferrite coatings on all conducting elements of the antenna increases film surface area, proving increased magnetic performance and full electrical antenna isolation.

The magnetic film of interest for our application is a spinel-type, oxide ferrite, and the deposition procedures for fabrication of such a film is no simple task, involving a myriad of deposition parameters in need of optimization. Formation of the sol, or precursor solution, requires careful stoichiometric calculations and measurements of required elemental percentages. Likewise, the heating temperature as well as mixing time and method play a large role in the quality of the resulting films. Several trials of sol composition were performed, varying the

percentages of included elements and chemicals, before the composition and sol preparation parameters were finalized.

The dip-coating apparatus (MTI Corp. PTL-MM01), shown in Figure 2.4, operates with a DC step motor and also has adjustable parameters that should be taken into consideration. The process works by inserting the sample into the solution until it is completely immersed, and then withdrawing it at a predefined, constant velocity. It is imperative that the withdrawal speed is constant and free of jitters in order to retain uniformity in the resulting film thickness. The significant variable dip-coating parameters include travel length as well as insertional and withdrawal speeds. The travel length is a trivial parameter to set, as this will just be the length of our antenna (10 mm) plus the distance between the solution surface and the bottom of the antenna when the sample is in its rest state. The withdrawal speed is a more complicated variable and must be optimized since this parameter effects the thickness of each film coating. Typically, the slower the withdrawal speed is set to, the thicker the resulting films will be.

FIG. 2.4. MTI Corp. PTL-MM01 dip-coating apparatus setup and components.

In order to achieve fully-encapsulating NiCo ferrite film, the sol-gel dip-coating deposition technique was employed with procedures modified from [25]. Prior to deposition, copper tape was applied to the areas of the antenna that require subsequent soldering, including the antenna transmission line, shorting pin location and ground attachment sites. Likewise, the antenna substrate was cleaned thoroughly beforehand with isopropyl alcohol in a magnetic stirrer. To begin preparation of the sol, precursor materials of iron (III), nickel and cobalt nitrates were weighed out in a stoichiometric ratio so as to achieve $\text{Ni}_{0.23}\text{Co}_{0.27}\text{Fe}_{2.50}\text{O}_4$ films. 50 g of iron

nitrate was dissolved in 100 ml of preheated (70 °C) ethylene glycol, while 6 g and 5.8 g of nickel and cobalt nitrates, respectively, were individually dissolved in 10 ml of preheated 2-methoxy-ethanol (2-MOE). Upon cooling, the three initial solutions were each mixed in an ultrasonic cleaner for 5 minutes. Finally, the precursor solutions were mixed together with a magnetic stirrer for 20 min, adding 10 ml of polyethylene glycol (PEG) for improvement of sol viscosity.

The antenna was then loaded into the dip-coating apparatus with an insertion and withdrawal speed of 70 mm/min. After each ensuing coating, the film was dried in air at 140 °C for 3.5 min followed by heating at 320 °C for 2.5 min in a tubular furnace. The ferrite solution tended to adhere only to the conducting elements of the antenna. After 10 coating cycles we performed, the substrate was annealed at 320 °C for 4 hours. Note that NiCo film was also annealed at 700 °C for 4 hours to validate the ferrite composition with clearer XRD patterns. After deposition onto the antenna substrate, the copper tape is removed to reveal uncoated soldering sites. Once the film fabrication is complete, characterization of the coatings can be performed to verify our ferrite composition as well as physical and magnetic properties.

2.3. Characterization Techniques

2.3.1. Antenna Characterization Methods

Several quantifiable metrics were used to characterize the fabricated implantable antenna, each requiring a unique instrument for measurement; measurements were made for the antenna in air and implanted 5 mm in the skin-mimicking gel. The first parameter examined is the reflection coefficient, which represents the reflection and transmission of wave frequencies in an electrical network. The corresponding reflection coefficient for the antennas were measured with

a vector network analyzer (VNA: Agilent N5230A), an instrument that measures the network parameters of the associated electrical networks. The reflection coefficient parameter is important to quantify because it identifies the device frequency of resonance and impedance matching at that frequency. A simple coaxial cable was used for connection of the antenna SMA port to the VNA.

Next, the power gain, or simply gain, of the fabricated antenna around its operation frequency is measured. The antenna gain is a significant figure of merit that considers both the antenna's directivity and electrical efficiency, and is calculated differently depending on the class of antenna. As a transmitter, the antenna gain reveals how efficiently the device converts input power into radio waves toward a specific direction. As a receiver, the antenna gain represents how efficiently the device transfers radio waves into electrical power when entering from a specific direction. For our application, the *frequency-dependent* realized gain is observed so that gain in and around the ISM band can be observed. Additionally, the observed angle is $(\theta, \phi) = (0, 0)$, which is of interest because it represents the direction normal to the antenna radiator, thus orthogonal to the human body when implanted. The gain of our antenna was measured with a combination of the VNA and an anechoic chamber (Raymond EMC QuietBox AVS 700). An anechoic chamber is a room or compartment designed to fully absorb reflections of either acoustical or electromagnetic waves, while remaining insulated from external noise sources. The purpose of this instrument is to simulate a noise-free space of infinite dimensions in order to characterize the antenna properties with little external signal interference.

Finally, the realized gain radiation patterns were observed at the antenna center frequency of resonance, and was characterized with the VNA and anechoic chamber. The gain radiation pattern of the antenna refers to the antenna's directional dependence of the strength of the radio

waves transmittance, in terms of the angular variation from the direction normal to the antenna's face; again, the observed angle is $(\theta, \phi) = (0, 0)$. This parameter effectively represents the directivity of the antenna, which is a metric that describes the antenna's radiating power density in its strongest direction, compared to the radiating power density by an ideal isotropic radiator (which emits uniformly in all directions) with equal radiated power. This metric is important because it reveals the direction of highest power radiation, which for our application should be in the direction normal to the antenna face, with radiated power attenuated in direction facing the body internal.

2.3.2. Ferrite Film Measurements

For characterization of the fabricated ferrite films, the films were first deposited on amorphous glass substrates, followed by deposition on the antenna substrate after material measurements. First, the thickness of the film was measured with a Veeco Dektak V220-Si surface profilometer. A profilometer is an instrument used to measure the profile of a surface, in order to quantify its thickness and roughness variations. The surface profiler for our measurements is a contact, or stylus, profilometer, which utilizes a diamond stylus that is put in contact with a sample and subsequently moved horizontally across the surface for a certain distance and contact force. This device can also be used to take magnified images of a sample. The target film thickness for our application will be elaborated on later in this work.

The most important metric to consider for our ferrite film will be the X-ray diffraction patterns, which were measured with a Philips X'Pert MRP X-ray diffraction apparatus utilizing Cu K_α radiation. XRD instruments provide conclusive atomic and molecular information regarding crystal structures and their atomic distances. This principle works by bombarding

monochromatic X-rays onto a crystalline material at a diffraction angle (θ), which are then scattered by the atoms in the sample. By varying the diffraction angle, the Bragg's Law conditions are satisfied:

$$2d\sin(\theta) = n\gamma , \quad (2.2)$$

where n is any integer, γ is the wavelength of the beam, and d is the spacing between diffracting planes (d-spacing). By plotting the angular positions and intensities of the resulting diffracted peaks, one can make quantifiable determinations as to the atomic and molecular structure of a sample, revealing its elemental composition. For our analysis, we hope to observe XRD peaks associated with NiCo spinel-type ferrite ($\text{Ni}_{0.23}\text{Co}_{0.27}\text{Fe}_{2.50}\text{O}_4$).

Finally, the magnetic properties, specifically the magnetic moment as a function of applied field, are measured with a Microsense EV9 vibrating sample magnetometer. The VSM instrument works by placing a sample inside a uniform magnetic field to magnetize the sample. The sample is then physically vibrated sinusoidally which induces a voltage in the system that is proportional to the sample's magnetic moment. The device varies the applied magnetic field in order to characterize the magnetic moment hysteresis loop as a function of this external applied field. The resulting magnetic moment that is observed is a measurement of the torque the material will experience in an external magnetic field. The two distinct types of VSM measurements we will consider are the *in-plane* and *out-of-plane* magnetic moment hysteresis loops.

The in-plane loop describes the magnetic properties for an applied field parallel to the surface plane of the film, while the out-of-plane measurements describe properties for an applied field orthogonal to the film surface. Both the in-plane and out-of-plane magnetic moment hysteresis loops are important because they describe the complete resulting magnetization

experienced in our ferrite material. We anticipate hysteresis measurements typical of soft magnetic materials, with relatively tall and narrow magnetic moment versus applied field loops. For our studies, the magnetic moment will be measured in standard electromagnetic units (emu) and the applied field in Oersted (Oe). Expectations are that the higher temperature annealed films will exhibit stronger magnetizations, however we require and hope to observe appropriately shaped hysteresis loops with magnetic saturation achieved for the films annealed at 320 °C as well.

CHAPTER 3

UNCOATED IMPLANTABLE ANTENNA RESULTS

3.1. Antenna Simulation & Optimization

We have already determined some crucial antenna design parameters to guide our optimization process. Rogers RO3010 laminate has been chosen as the substrate with a thickness of 1.28 mm and antenna surface dimensions of $10 \times 10 \text{ mm}^2$. Additionally, the desired frequency band of operation is the ISM band (2.4 – 2.5 GHz). Typically, microstrip patch antennas are fed by a microstrip transmission line (Fig. 1.1), and this is chosen as the preferred feed technique for our implantable antenna application. This feed technique is chosen due to simplicity and ease of soldering attachment to coaxial connectors. With these constraints in place, antenna simulation optimizations can be performed to produce antenna resonance in the desired band with efficient antenna performance parameters.

To begin our radiator optimization process, we used a typical *rectangular* microstrip patch antenna model as our initial framework to build on (Fig. 1.1). One of the more typical radiator design methods to aid in antenna miniaturization is extending the path of current flow via a longer radiator surface conducting channel; example radiator geometries that incorporate this principle are serpentine, spiral and meandering, shown in Fig. 3.1. It can be seen that, as opposed to the traditional rectangular patch, the flow of current is extended with these more complex radiator geometries. In an attempt to reduce the antenna frequency to just above the ISM band, we will incorporate this current path lengthening principle into our antenna radiator design, as well as other antenna miniaturization methods.

FIG. 3.1. Typical (a) serpentine, (b) spiral and (c) meandering antenna radiator geometries [2].

Initial simulations were performed for variations of the *rectangular* and *spiral* radiator geometries. The results from these simulations yielded decent antenna performance parameters, but did not exhibit miniaturization beyond that in current literature. Therefore, upon further simulations with a *loop* radiator geometry, we opted to combine these designs into a single radiator configuration; this is the first work to incorporate these various geometries into a combined design. The primary variables in the patch optimization process include spiral arm dimensions, transmission line width as well as shorting pin radius and location. All simulations were of the discrete, linear-step type, with a 10 MHz step size. It is also worth noting that though air simulations are performed for subsequent antenna validation, the design of the implantable antenna is performed with the device embedded 5 mm in the human tissue phantoms, since the human body will be its ultimate operation destination.

Our preliminary implantable antenna design is depicted in Fig. 3.2, implanted 6.4 mm into the three-layer tissue phantom. The initial radiator design is composed of a novel geometry that we call the *asymmetric spiral-loop* geometry. The radiator consists of an asymmetric spiral that ultimately coalesces into a loop, with its right side resembling a traditional rectangular patch.

This configuration aids in extending the current path, and, as a result, reducing the antenna resonance frequency. We also incorporated a copper shorting pin (radius = 0.3 mm) that couples the antenna radiator and ground plane ($10 \times 10 \text{ mm}^2$), producing a PIFA. The reason for retaining a partially rectangular radiator geometry is because parametric simulations for shorting pin location revealed best antenna miniaturization when radiator and ground plane coupling is at a point with larger radiator surface area. In this configuration, the optimal shorting pin location is found to be near the lower right corner of the antenna radiator.

FIG. 3.2. Preliminary implantable antenna design with asymmetric spiral-loop radiator geometry.

In Figure 3.2, the fundamental elements of the initial implantable antenna design are listed. The feed strip is a required feature for HFSS simulation, designating the location where current is ultimately collected from the antenna, found at the end of the radiator transmission line. The unique radiator geometry can be observed, where the spiral portion of the radiator stems from the top of the rectangular section and, after three turns, loops into itself. All of the conducting elements of the antenna are simulated as copper (or perfect E), which is the same material as the cladding on the RO3010 laminate. Also in this design is the addition of upper and lower *superstrates* (upper not shown), which is another commonly used antenna miniaturization technique. The idea is that shielding the antenna with high permittivity dielectric materials, such as Rogers materials, assists in antenna miniaturization because they shorten the effective antenna wavelength, resulting in lower resonance frequencies. However, inclusion of thicker superstrate layers ultimately produces larger antenna dimensions. The antenna reflection coefficient, representing effective impedance matching, is shown in Fig. 3.3 and reveals the antenna frequencies of resonance.

FIG 3.3. HFSS simulated reflection coefficient for preliminary implantable antenna design in the three-layer tissue phantom.

The reflection coefficient plot for the preliminary antenna design in the simulated tissue model achieves resonance at 1.59 and 3.53 GHz with return losses of 3.58 and 13.48 dB, respectively. Obviously, a primary resonant frequency of 3.53 GHz is too high for our ISM band application, requiring design modifications to reduce this by about 1 GHz. However, the inclusion of this secondary resonance (1.59 GHz) can be potentially useful with application as an activation band, used to initiate, or “turn-on”, the device for power and data transmission within the primary (ISM) band [26]. The associated three-dimensional (3D) antenna realized gain radiation patterns at each resonance are shown in Fig. 3.4.

FIG. 3.4. HFSS simulated realized gain 3D radiation patterns for preliminary implantable antenna resonances.

In Figure 3.4, the peak realized gains for resonances at 1.59 and 3.53 GHz are shown to be -18.4 and -16 dBi, respectively, with energy propagation concentrated at the antenna plane facing away from the body internal. These values are consistent with required gain levels for implantable antenna applications; during our optimization process we will attempt to keep the realized gain for the primary resonance above -18 dBi. Likewise, in the subsequent antenna optimizations we will attempt to retain these radiation configurations with strong wave propagation in the direction away from the body and attenuation of energy absorption towards the body internal. With an understanding of the required property adjustments, antenna modifications were performed in an attempt to lower the primary antenna resonance while retaining appropriate antenna gain levels and the secondary resonance.

First, the decision was made to remove the antenna superstrate layers because the associated miniaturization benefits were not significant enough to compensate for the nearly triple increase in total antenna volume. Additionally, these layers will theoretically not be needed for insulation of the antenna conducting elements, as it is the hope that subsequent ferrite

deposition will serve this purpose. After the removal of these layers, we used HFSS parametric simulations to perform various dimensional optimizations to our preliminary asymmetric spiral-loop radiator geometry. These parametric simulations include dimensional variations for the microstrip transmission line and spiral/loop radiator components.

Foremost, the resonant frequency of the preliminary antenna model is approximately 1 GHz greater than the f_c our desired frequency band of operation. Varying the width of the spiral/loop radiator components is shown to have a significant effect on resonance, and each component was optimized accordingly to produce resonance just above the ISM band. Similarly, the shorting pin location and dimensions are pivotal in defining resonance, and is shown to produce larger surface current when located near the corners of the rectangular radiator portion. The optimal distance between edges of the antenna substrate and the edges of the radiator were also investigated because, typically, an increase in overall radiator surface area produces generally lower antenna resonant frequencies. There are of course exceptions to this statement, however, it must be kept in mind that we are not looking to simply reduce the resonance as much as possible, but to reduce the operation frequency to just above the ISM band. Upon completion of various parametric simulations, we have determined the layout and radiator dimensions for our optimized implantable antenna design, shown in Fig. 3.5.

FIG. 3.5. Layout (a) and patch dimensions (b) of the optimized implantable antenna design.

The final, optimized implantable antenna design has overall dimensions of $10 \times 10 \times 1.28$ mm (volume = 1.28 mm^3), with a shoring pin radius of 0.6 mm and a transmission line width of 0.5 mm. These optimized parameters yield appropriate antenna resonances and reflection

coefficients for our ISM band application for the antenna simulated while implanted 5 mm into the human skin model, as seen in Fig. 3.6. The simulated antenna implanted 5 mm in human skin has a primary resonance at 2.55 GHz with a -11 dB reflection coefficient, as well as a secondary, or “turn-on”, resonance at 1.185 GHz with a -12 dB reflection coefficient. The primary resonant frequency is 100 MHz above the center of the ISM band (2.45 GHz), which, as desired, allows space for the reduction of this resonance into the band via ferrite film deposition. Additionally, the implanted antenna has a -10 dB bandwidth of 60 MHz at the primary resonance.

When the antenna is simulated in air, the resulting frequencies of resonance are increased slightly and the associated reflection coefficients are diminished. This phenomenon occurs because the increased permittivity of the surrounding human tissue provides for increased impedance matching and shortening of the effective antenna wavelength over an air environment. It is also expected that the antenna operating in air does not meet our required return loss and gain specifications, because the antenna design and optimization process was employed for the antenna implanted in human tissue phantoms. The resulting air parameters will be used solely for subsequent antenna performance validation.

FIG. 3.6. HFSS simulated reflection coefficient for optimized implantable antenna in air and human skin phantom.

The frequency-dependent realized gain realized gain for the implantable antenna in air and skin-mimicking phantom in the range of 2 – 3 GHz is shown in Fig. 3.7, revealing gain parameters in and around the primary antenna resonance. Simulated realized gain values of -17 and -19 dBi were observed at the primary resonance for the antenna in skin and air, respectively. Similarly, the gain (G) variance at ISM band limits, $\Delta G_{f(2.5\text{GHz})-f(2.4\text{GHz})}$, is 1.97 dBi. These values are appropriate for our application, with hopes that the realized gain at resonance can be improved further with subsequent ferrite deposition. As expected, the realized gain at resonance for the antenna in air is slightly improved over the implanted antenna due to the higher dielectric loss tangent associated with human tissues. This illustrates the performance deteriorations that

are caused by antenna tissue envelopment, and the challenges associated with implantable antenna design.

FIG 3.7. HFSS simulated frequency-dependent realized gain $(\theta, \phi) = (0, 0)$ for optimized implantable antenna in air and human skin phantom.

3.2. Uncoated Antenna Measurements

FIG 3.8. Fabricated implantable antenna.

The fabricated, uncoated implantable antenna is shown in Fig. 3.8. Characterization of the fabricated implantable antenna was performed in air and tissue gel environments in order to validate our expected simulated parameters. The experimental measurement setup is shown in Figure 3.9, with the antenna embedded 5 mm in the 12 mm thick skin-mimicking gel; this implantation environment is analogous to our HFSS simulated human skin phantom. The reflection coefficient of the fabricated antennas in air and implanted in the skin-mimicking gel were characterized with the VNA.

FIG 3.9. Antenna measurement setup with VNA and skin-mimicking gel.

The measured and simulated reflection coefficients of the designed antenna in air and human skin-mimicking gel are shown in Fig. 3.10. The fabricated implanted antenna resonates at 2.56 GHz with a measured reflection coefficient of -16 dB and a 180 MHz -10 dB bandwidth. In air the antenna resonates at 2.88 GHz, with a return loss of less than 2 dBi. Minor discrepancies between measured and simulated data are observed for the implanted antenna, which are most likely attributed to the increased dielectric loss of our fabricated skin-gels over values of human skin used in simulations. The implanted antenna has a secondary resonance at 1.2 GHz (return loss of 6 dB) with potential application as an activation band, used to initiate, or “turn-on”, the device for power and data transmission within the primary (ISM) band. Note that the simulated and measured reflection coefficients for the antenna characterized in air are in excellent

agreement, which validates that the fabricated antenna is a good representation of our simulated model.

FIG 3.10. Measured and simulated reflection coefficients for antenna in air and implanted in human skin-mimicking gel.

The measured and simulated realized gain $(\theta, \phi) = (0, 0)$ for the designed implantable antenna in air and skin-mimicking gel in the range of 2 – 3 GHz are shown in Fig. 3.11. The measured RG of -17 dBi at resonance, when the antenna is embedded in the skin-mimicking gel, is slightly lower than that of the antenna in air ($RG = -15$ dBi) due to the associated tissue losses. Again, the peak realized gain values for the simulated and measured antenna in air are in very good agreement. The gain variance at ISM band limits, $\Delta G_{f(2.5\text{GHz})-f(2.4\text{GHz})}$, is 3.9 dBi, slightly higher than the simulated value (1.97 dBi). It should be noted that the two simulated implant

conditions (skin and three-layer phantoms) yield highly similar resonances and reflection coefficients for the antenna in Fig. 3.6 and Fig. 3.10, as well as comparable ISM band gain spectra in Fig. 3.7 and Fig. 3.11. Overall, the simulated and measured implantable antenna performance parameters are in good agreement, allowing subsequent comparisons to be made between the bare and ferrite-coated antenna variations.

FIG 3.11. Measured and simulated frequency-dependent realized gain $(\theta, \phi) = (0, 0)$ for antenna in air and implanted in human skin-mimicking gel.

CHAPTER 4

IMPLANTABLE FERRITE ANTENNA RESULTS

4.1. Coated Antenna Simulations

The antenna radiator dimensions and shorting pin location are optimized to achieve the best impedance matching at a resonance just above the ISM band, given our spatial constraints. In order to further miniaturize and improve antenna performance, we have performed parametric simulations with the implantable antenna coated with ferrite films of varying material parameters. The most significant parameters to consider when choosing the desired ferrite film configuration include film thickness, relative permeability, and magnetic tangent loss. For our implantable application, we desire a film with relatively high permeability and low magnetic loss. A high magnetic permeability is sought because the property describes the ability of a material to support the formation of a magnetic field within itself, which in turn aids in wave propagation, and thus directivity of our antenna. Similarly, the antenna quality factor (Q), which characterizes antenna bandwidth relative to its center frequency, decreases with the addition of increasing magnetic loss from the ferrite film [27]. A larger Q signifies a reduction in the rate of energy loss relative to the reserve energy of the radiator and, therefore, the oscillations die out more slowly. Typically, magnetic materials exhibiting high permeability and low loss are described as magnetically *soft* materials.

With typical soft ferrite materials, as the frequency of operation increases, the permeability decreases and the magnetic loss increases. Therefore, it is quite difficult to fabricate ferrite films exhibiting our desired magnetic properties at relatively high (GHz) operation

frequencies, especially at reduced deposition temperatures. However, in [9], a $\text{Ni}_{0.23}\text{Co}_{0.27}\text{Fe}_{2.50}\text{O}_4$ film is reported exhibiting a relatively high μ_r and low $\tan \delta_\mu$ of 10 and 0.3, respectively, at 2.45 GHz. Furthermore, nickel-cobalt films are reported to have a relatively low electrical conductance [10] and toxicity [11]. The reported film was fabricated via the sol-gel dip-coating deposition technique. For our ferrite antenna simulations, we modeled our film based on the properties of this reported NiCo film. In addition, parametric simulations were performed to observe the effect of ferrite film thickness (thickness = 5 and 10 μm) and magnetic loss ($\tan \delta_\mu = 0.05, 0.1$ and 0.3) with a thickness of 10 μm on antenna performance.

The geometric configurations of the simulated ferrite films are such that they shroud all of the conducting elements of the antenna, including the antenna ground plane and radiator, with a uniform thickness. Beyond the addition of ferrite films, all other environmental, boundary, simulation and implantation conditions remain consistent between the ferrite-coated and non-coated antenna simulations in human tissue phantoms. The first film parameter to be varied is film thickness, and the corresponding effect on antenna reflection coefficient will be investigated. For the film thickness simulations, comparisons were made for the antenna with no ferrite, as well as with films of 5 and 10 μm thickness, each with a μ_r and $\tan \delta_\mu$ of 10 and 0.3, respectively. The resulting simulated effect of ferrite film thickness on reflection coefficient for the antenna implanted in three-layer tissue phantom is depicted in Fig. 4.1.

FIG 4.1. Effect of ferrite film thickness on reflection coefficient for antenna implanted in human skin tissue phantom.

From Figure 4.1, inclusion of ferrite film coatings results in reduced resonant frequencies as well as improved reflection coefficients. The addition of 10 μm thick magnetic layers have effectively lowered the frequency of resonance by 70 MHz, with a reduced center frequency (2.46 GHz) near the center of the ISM band, as desired. The 10 μm thick films also reveal improvements in resonant return loss and -10 dB bandwidth by 3 dB and 40 MHz, respectively. It should be noted that miniaturization effects begin to saturate at film thicknesses above 10 – 15 μm , where the additional volume is no longer compensated, or justified, by the noticeable antenna performance enhancements. Therefore, this will represent our desired thickness range (10 – 15 μm) for subsequent ferrite film deposition. Next, the effect of varying film thickness and magnetic loss tangent on frequency dependent realized gain is observed.

Figure 4.2 reveals the effect of varying ferrite film thickness and magnetic loss on frequency dependent realized gain for the antenna implanted in the human skin tissue phantom. It is found that ferrite film coating improves realized gain (RG) within the ISM band, with largest gain increases at resonance found for films having lower magnetic loss. The RG can be expressed as:

$$RG = RE \cdot D \cdot (1 - |\Gamma|^2), \quad (4.1)$$

where RE is the radiation efficiency, D is the directivity, and Γ is the reflection coefficient.

Accordingly, the RG was improved because of higher directivity and better impedance matching by loading ferrite film [27–28]. Additionally, the gain variance at ISM band limits, $\Delta G_{f(2.5\text{GHz})-f(2.4\text{GHz})}$, decreased from 1.97 to 0.44 dBi for the antenna with 10 μm films over the non-ferrite antenna. This suggests that a more reliable wireless signal in the operating frequency band can be achieved with ferrite loading. Moreover, the observed antenna performance enhancements are attributed to the robust demagnetization field of magnetic thin films, which allow for GHz ferromagnetic resonance (FMR) frequencies [7–8]. It has been found that the desired ferrite film parameters for our application include a film thickness between 10 – 15 μm and as low a magnetic loss tangent as possible. With preliminary implantable ferrite antenna design and simulation complete, fabrication and characterization of the uncoated antenna can begin.

FIG 4.2. Effect of varying film thickness and magnetic loss on realized gain ($\theta, \phi = (0, 0)$) for antenna implanted in human skin tissue phantom.

4.2. Ferrite Film Properties

The ferrite film-coated implantable antenna is depicted in Fig. 4.3. It can be seen that the magnetic film tended to adhere only to the conducting surfaces of the antenna (i.e. radiator and ground plane); however, the coatings do fully encapsulate these conducting elements, aiding in biocompatibility of the device. The SMA cable was modified and connected in the same manner as the uncoated antenna. The ferrite films deposited on amorphous glass substrates were characterized with several metrics including XRD, VSM and surface profilometry analysis. The film thickness was measured with a Veeco Dektak V220-Si surface profilometer to be approximately 15 μm, making our film thickness for one coating iteration approximately 1.2 μm.

The measured film thickness value is within our desired range of 10 – 15 μm . Additionally, the film exhibited a relatively smooth surface, with little thickness variation throughout the sample.

FIG 4.3. Fabricated ferrite film-coated implantable antenna.

XRD analysis was performed with a Philips X'Pert MRP X-ray diffraction apparatus using Cu K_{α} radiation. The desired peaks for this application are that of spinel type, oxide, nickel-cobalt ferrite. Based on the work by prior authors [9], the primary peaks we expect for our ferrite films are [220], [311], [400] and [440], with corresponding 2θ angles of approximately 30° , 36° , 43° and 62° , respectively. All of these peaks are observed in the resulting XRD patterns for films annealed for 4 hours at 320°C and 700°C , shown in Fig. 4.4. As expected, the film annealed at the higher temperature generally produces more defined peaks at the points of interest, though the nickel-cobalt ferrite composition is confirmed in films for both annealing temperatures. This means that both films should have magnetic properties corresponding to

spinel type, oxide, nickel-cobalt ferrite, with higher temperature annealed films exhibiting stronger properties. In order to investigate these properties, magnetic moment measurements for the films are observed.

FIG 4.4. XRD pattern for deposited NiCo oxide spinel-type ferrite film annealed at (a) 320 °C for 4 hours and (b) 700 °C for 4 hours.

Vibrating sample magnetometer analysis was performed with a Microsense EV9 VSM apparatus for both in-plane and out-of-plane magnetization orientations, shown in Fig. 4.5. Hysteresis loops with low coercivity are observed in films annealed for 4 hours at both 320 °C and 700 °C. Both samples of different annealing temperatures achieve positive saturation around the same applied field strength of 1000 Oe, and hysteresis loops for both samples are symmetric about the zero applied field axis. These relatively tall and narrow hysteresis loops are typical for soft magnetic materials, which is what we expect for our nickel-cobalt ferrite films. As anticipated, the film annealed at the higher temperature produces a more stable hysteresis loop, exhibiting higher saturation and remnant magnetization.

The film annealed at 320 °C achieves a positive, in-plane saturation of approximately 0.003 emu, while the film annealed at 700 °C saturates around 0.2 emu. Likewise, the film annealed at 320 °C achieves a positive, in-plane remnant magnetization of approximately 0.001 emu, and the film annealed at 700 °C has a remnant magnetization around 0.1 emu. In both samples, the in-plane magnetization is slightly higher than the out-of-plane magnetization; though the differences are small, this indicates that the dominant plane of magnetization is in the direction parallel to the surface of the film. Though the higher temperature annealed films exhibited stronger magnetic properties, the films annealed at 320 °C still exhibit soft magnetic material properties acceptable for our application. However, these film characterization results illustrate the obstacles experienced in low-temperature ferrite deposition. Due to the issues associated with process, other low-temperature ferrite fabrication methods should be investigated more thoroughly in future works.

FIG 4.5. In-plane and out-of-plane VSM measurements for NiCo ferrite films annealed at (a) 320 °C for 4 hours and (b) 700 °C for 4 hours.

4.3. Ferrite Antenna Measurements

The reflection coefficient of the fabricated antennas in air and implanted in the skin-mimicking gel were characterized with the VNA. With regard to radiation characteristics of the fabricated antennas, an anechoic chamber (Raymond EMC QuietBox AVS 700) was used, along with the VNA. The addition of dip-coated NiCo ferrite coatings effectively reduces the resonant frequency by 110 MHz to the center of the ISM band (2.45 GHz), with a 4.2 dB return loss

improvement at resonance as presented in Fig. 4.6. The observed resonant frequency and return loss improvements are of desired values for operation in our chosen frequency band, with ferrite coating benefits consistent with expectations from simulations. The ferrite antenna also reveals a 35 MHz increase in -10 dB bandwidth over the uncoated antenna. The improved bandwidth of 215 MHz now covers the complete ISM spectrum of interest.

FIG 4.6. Measured reflection coefficient comparison for antennas implanted in human skin-mimicking gel.

In addition to resonant frequency and return loss comparisons, measurements are also investigated to observe the benefits of ferrite loading on antenna realized gain. Figure 4.7 illustrates the frequency-dependent RG comparison, with the coated antenna experiencing a 6 dBi average gain increase within the ISM band limits, resulting in a peak RG of -17 dBi.

Additionally, the gain variance, $\Delta G_{f(2.5\text{GHz})-f(2.4\text{GHz})}$, decreased from 3.9 to 0.1 dBi for the coated antenna over the non-ferrite antenna, revealing an extremely stabilized gain within the ISM band limits. This gain stabilization refers to the realized gain spectrum becoming more horizontal, i.e. less variation in magnitude, over the frequency band of interest. This is perhaps the most striking and significant benefit observed from our ferrite loading experiments, because having a stable gain over an antenna's frequency band of operation is imperative to the performance and efficiency of any device utilizing wireless communication.

FIG 4.7. Measured frequency-dependent realized gain $(\theta, \phi) = (0, 0)$ comparison for antennas implanted in human skin-mimicking gel.

The normalized realized gain radiation patterns at the elevation plane for the antennas at 2.5 GHz are compared in Fig. 4.8. The overall radiation geometry of the uncoated antenna is

retained in the ferrite loaded antenna, with one significant difference observed. For the coated antenna, it can be seen that the direction of greatest wave propagation is in the direction facing away from the body, where a corresponding, external transmitting antenna would theoretically be located. Likewise, wave propagation is attenuated in the antenna direction facing the body's internal (by over 10 dBi), effectively improving antenna directivity and reducing the energy absorbed by the tissues surrounding the device.

These radiation principles are particularly significant in implantable device applications, due to the regulations imposed by the FCC limiting the allowable amount of electromagnetic energy absorbed by human tissue from external electronic devices. Since an external transmitting antenna would be placed facing out of the body, there is no benefit in having a large antenna gain in the direction facing away from this energy source. Therefore, the observed gain attenuation will aid in biocompatibility of the antenna and allow compliance with the strict regulations imposed on implantable medical devices. The ferrite antenna results reveal significant improvements over the uncoated antenna, indicating implantable antenna miniaturization and reliable wireless power transfer and data communication in the operating frequency band can be realized with ferrite loading.

FIG 4.8. Measured normalized antenna gain radiation pattern comparison.

CHAPTER 5

CONCLUSION

We have developed a miniaturized implantable ferrite antenna operating in the ISM band, with a -20 dB reflection coefficient and -17 dBi realized gain at resonance (2.45 GHz). The addition of dip-coated ferrite film decreases the resonant frequency by 110 MHz with a 4.2 dB return loss improvement at resonance and 35 MHz increase in -10 dB bandwidth. Ferrite loading has also shown to improve realized gain by an average of 6 dBi in the antenna frequency band of operation, while also improving directivity over the uncoated antenna. Most strikingly, the ISM band gain variance, $\Delta G_{f(2.5\text{GHz})-f(2.4\text{GHz})}$, decreased from 3.9 to 0.1 dBi for the coated antenna over the non-ferrite antenna, revealing strong gain stabilization within the ISM band limits.

Extensive simulations and experimental validations were performed for the fabricated ferrite and non-magnetic antennas. Implantable antenna characterization was performed with the device embedded in high quality skin-mimicking gels with dielectric properties analogous to those of human skin. Experimental antenna results are in good agreement with simulations, validating the design and efficacy of our fabricated implantable antenna. Parametric simulations were also conducted to observe the effects of varying ferrite film parameters on antenna performance. It was found that significant performance benefits were observed for films with higher relative permeability and lower magnetic tangent loss. Furthermore, best antenna improvements were observed for ferrite film coating between 10 – 15 μm , with no further antenna performance enhancements observed above this thickness.

These results indicate that ferrite film loading can aid in implantable antenna performance and miniaturization. Additionally, ferrite films are non-toxic and electrically insulating while maintaining high magnetic properties up to several gigahertz, making them excellent candidates for radio-frequency biomedical applications. With evidence suggesting that the inclusion of magnetic materials can produce a myriad of benefits in implantable antenna systems, further work is required in the investigation of more efficient, low-cost, low-temperature ferrite film deposition techniques. Upon further validation and understanding of the enhancements resulting from magnetic material incorporation, the principles outlined in this work have the potential for application to a wide range of RF devices and systems.

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