Wireless RF Induced Energy Absorption and Heating of Lanthanum-Nickel Alloy in the Near-Field

Michael Dillon Lindsay

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WIRELESS RF INDUCED ENERGY ABSORPTION AND HEATING OF LANTHANUM-NICKEL ALLOY IN THE NEAR-FIELD

by

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Bachelor of Science
University South Carolina, 2018

Submitted in Partial Fulfillment of the Requirements
For the Degree of Master of Science in
Electrical Engineering
College of Engineering and Computing
University of South Carolina
2019

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Mohammad Ali, Director of Thesis
Guoan Wang, Reader
Cheryl L. Addy, Vice Provost and Dean of the Graduate School
DEDICATION

To my wife, Anna. You have provided an unmeasurable amount of support throughout this process. This work would not be possible without your motivation and love.
ACKNOWLEDGEMENTS

I am extremely appreciative of Dr. Mohammod Ali for the time he has always created for me and this work. I endeavor to surround myself with people of his caliber, dedication, and humbleness.

I am thankful for Michael Brown and his team at the Savannah River National Laboratory. His acute methodologies are the support behind all the experimental validation stages.

I am thankful for Jeremey Gilliam, Kyle George, and Bill Bradley at the University’s Machine Shop. Without their quality work, the accuracy and precision in our measurements would suffer.

I am thankful for David Metts. His true friendship provided many meaningful conversations with a cup of coffee.

Lastly, I am thankful for my family. Without their support, I would not have been able to start my education at such a prestigious university.

May this work glorify God and reflect the fortune and blessings I have received and experienced from every individual I have encounter during my academic career.
ABSTRACT

This thesis will investigate whether and how a synthetic metal-magnetic alloy absorbs enough near-field electromagnetic (EM) radiation for heating. Although resistive elements could be used for indirect heating, a wireless radio frequency (RF) method is explored because of its advantages in its non-intrusive nature and its ability to direct or focus energy within a specific area. For our case, the RF heating process relies on the materials ability to absorb sufficient RF energy due to the induced currents. We expect there to be significant surface resistance due to the conductivity and magnetic permeability of the material and thus heat. This ratio between these unknown parameters ($\sigma/\mu_r$) are determined to be $1 \times 10^6$ with surface roughness and $2.5 \times 10^5$ without surface roughness.

When using a resonant dipole at 2.4 GHz, not all the RF energy absorbed by the material will become heat so an Efficiency Factor (EF) was created to normalize the simulated results to the experimental verification using a well-known material, graphite. The EF was determined to be 1.77% and could be used for the unknown alloy calculations, along with a Spacing Factor (SF) of 14.7% to account for the granular material. For graphite at 5/8 inch from 10 W input power antenna, the temperature rise after 5 minutes was 7.9 °C. Similarly, for the granular alloy the temperature rise was 4.5 °C. A new method of further enhancing the EM energy absorption in material was developed by designing and investigating a split ring resonator (SRR) in simulations which showed the potential to increase the temperature rise in the alloy by a factor of 5.5 due to its inherently resonant nature.
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CHAPTER 1. INTRODUCTION

RF heating has been an interesting subject since the increasing research in Radar technology during World War II. Since then, the two largest uses are seen in the food and medical industries [1]. Raytheon engineers were able to design and patent a product that could quickly heat meals by means of dielectric heating. This device was first named the “Radarange” because of its utilization of radio frequency heating and is known today as the microwave oven [2]. Microwave technology continues to grow in the medical field with the same operating principle and desired outcome: close range heating with high frequency EM radiation. Hyperthermia treatment and ablation techniques are used with biological tissue to kill and eliminate harmful cells in the human body with heat [1]. These two examples, along with many other common uses, use dielectric heating that take advantage of the water molecules that are present in the materials. Because of its polar structure, the water molecules will rapidly rotate to align with an applied alternating field, creating friction along with heat.

The focus of this thesis is to investigate and analyze whether and how a synthetic alloy consisting of part metal (Lanthanum) and part magnetic (Nickel) materials can absorb enough electromagnetic energy from the near fields of an antenna and thus be effectively heated. Traditionally, such an alloy can be heated using resistive heating where resistive wires carrying currents cause the heating. Heating using a wireless RF method of a metal-magnetic alloy is rather unusual since such materials cannot be heated using the traditional microwave method of dielectric heating. Although it is not common, this method has
significant advantages which include its non-intrusive nature and its ability to direct or focus energy within a specific area.

For our case with the metal-magnetic alloy, the RF heating process must rely on its ability to absorb sufficient RF energy due to the induced currents flowing in the material. For perfect conductors, there will be no RF energy absorbed in the material. For a metal-magnetic alloy that is made of granular elements, we expect there to be significant surface resistance due to the conductivity and magnetic permeability of the material. This surface resistance is expected to cause losses in the material and thus heat.

Unfortunately, no information is available about the constitutive parameters of this alloy, such as its electrical conductivity, magnetic permeability, and surface roughness. Therefore, firstly, the focus of this thesis is to investigate and develop a method to understand and define these properties. A transmission line based simulation and experimental method will be developed and the results will present these characteristics. Secondly, the effectiveness of RF energy absorption in the material due to a log periodic dipole antenna (LPDA) is studied, which show some challenges with the approach. Thirdly, this thesis focuses on studying the RF absorption in the material due to a dipole antenna source. To accurately calibrate the RF heating from an antenna source in the simulation model, such as, what fraction of RF energy is actually being used to heat the material, a substance with known electrical property (graphite) is experimentally used and then compared with simulation. Finally, the shape of the absorbing material is investigated to determine whether the proposed granular alloy material can be formed into a frequency selective surface comprised of corrugated split ring resonators and heated more effectively than its random layout.
CHAPTER 2. RF HEATING AND FREQUENCY SELECTIVE SURFACE BACKGROUND

Hyperthermia treatment and ablation techniques are among the most popular uses of RF heating for organic tissue. Hyperthermia and ablations are used to kill and eliminate harmful cells in the human body. These two procedures together outline the fundamental concepts with wireless power transfer into an experimentally measured medium. RF heating of inorganic materials also have classic usage examples, with the main cases being dielectric welding and induction heating. Because the techniques are well documented and understood, a literature review is presented to cover the basic terminology with each topic, the theory of operation in each case, and the current devices that are utilized.

2.1. RF HEATING OF ORGANIC TISSUES

Heating organic tissues with RF energy was a safety concern as handheld communication devices became more popular. Today, the interactions with human tissues and EM fields are better understood and higher resolution head and body models are readily available for simulations with newer devices.

2.1.1. SKIN DEPTH OF TISSUES

Electromagnetic radiation has an effective depth that it can penetrate into a material. This skin depth is a function of the material’s properties and is defined as the distance from the surface of a material inward at which the induced currents from the EM radiation fall to \( e^{-1} \) (36.8%) of the value at the surface [1].
It is a concern that energy radiated from an antenna could cause harm to biological tissue. Because of the skin effect, this concern deals with the boundary of biological tissue closest to the radiating element. For antennas outside of the body, this boundary is the human skin and the muscle tissues just under the skin surface, but interstitial antennas will have this effect on tissues inside the body near the radiating element.

Table 2.1 is extracted from [4] and shows the skin depths in human tissue when exposed to different EM radiation sources outside of the body. This skin depth is calculated using (2.1), given that the resistivity of the human skin is 35.3 Ω cm when considering the parallel combination of tissues just under the skin [4], [5].

Table 2.1 Typical Skin Depths in Human Tissue, Extracted from [4]

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Radio FM</th>
<th>TV Transmitter</th>
<th>Telephony Mobile</th>
<th>Telephony Mobile</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency (MHz)</td>
<td>100</td>
<td>450</td>
<td>900</td>
<td>1800</td>
</tr>
<tr>
<td>Skin depth (cm)</td>
<td>3</td>
<td>1.5</td>
<td>1</td>
<td>0.7</td>
</tr>
<tr>
<td>Depth at which power reduces to 1% (cm)</td>
<td>9</td>
<td>4.5</td>
<td>3</td>
<td>2</td>
</tr>
</tbody>
</table>

\[ \delta = \frac{2}{\omega \mu \sigma} = \frac{2\rho}{\omega \mu} \quad (2.1) \]

Figure 2.1 graphically shows how the absorbed power reduces as a function of penetration depth in the skin with individual traces ranging from 10 – 2450 MHz.

Biological tissue is said to be dispersive because the relative permittivity varies as a function of frequency and can be modeled many different ways. Figure 2.2 shows this relationship. This figure is plotted logarithmically to convey the shape of the negative trend. As frequency increases, the relative permittivity of living material begins to approach 80.2, the relative permittivity of water at room temperature [6].
Figure 2.1 Power absorbed in muscles as a function of the skin depth at various frequencies. Extracted from [4]

Figure 2.2. Dielectric constant of living material as a function of frequency. Extracted from [6]
2.1.2. Specific Absorption Rate

In order to determine the effects of EM radiation absorbed by biological tissues, the specific absorption rate (SAR) is used to measure the rate at which a mass absorbs energy [7]. This measurement is not unique to biological tissue and it can be used to quantify any such measurement, however it mostly stays within the RF field. The rate of energy over time absorbed by a mass can be described as the power absorbed by the mass and is usually presented in the following units [8], [9].

\[
\text{SAR} = \left[ \text{Joule/sec/kg} \right] = \left[ \text{Watt/kg} \right]
\]

This value can be calculated for any volume with known conductivity, electric field, and mass density. All three of these values can change as a function within the volume and are described in (2.2)

\[
\text{SAR} = \frac{1}{V} \int \frac{\sigma(x) |E(x)|^2}{\rho(x)} dx
\]  

(2.2)

\(V\) is the volume of concern, \(\sigma(x)\) is conductivity of the volume as a function of \(x\), \(E(x)\) is electric field vector in the volume as a function of \(x\), and \(\rho(x)\) is the density of the volume as a function of \(x\) [9].

If the parameters do not change within the defined volume, it is simpler to use one of the variations of this equation.

\[
\text{SAR} = \frac{\sigma E_{\text{rms}}^2}{\rho}
\]  

(2.3)

\[
\text{SAR} = \frac{\omega \varepsilon_0 \varepsilon_r E_{\text{rms}}^2}{\rho}
\]  

(2.4)
It is important to note the dependence of SAR on the frequency of the radiating fields, as highlighted in the above equation, however, this depends on parameter dispersion.

The SAR is useful because it can be used to model the temperature change in tissue. Temperature increases in living cells is the largest concern with electromagnetic radiation. (2.6) is used to determine the temperature change in human tissue as follows [7].

\[
\frac{\Delta T}{\Delta t} = \frac{\text{SAR} + P_m - P_c - P_b}{C}
\]  

(2.6)

In this equation \(\Delta T\) is the change in temperature, \(\Delta t\) is the exposure time, \(P_m\) is the metabolic heart rate, \(P_c\) is the rate of heat loss from thermal conduction, \(P_b\) is the rate of heat loss from blood flow, and \(C\) is the specific heat capacity. It is also important to note that before the exposure, the metabolic heart rate is equal to the sum of the rate of other heat losses. This means that the equation can be simplified to the following during initial conditions or relatively short exposure times, usually less than an hour [7].

\[
\frac{\Delta T}{\Delta t} = \frac{\text{SAR}}{C}
\]  

(2.5)

From equation (2.5) and Table 2.2, which was extracted from [7], the SAR can be experimentally derived in any of the listed human tissues. For example, if a selected volume of fat tissue was exposed to EM radiation for 30 minutes and it was recorded to increase temperature by 1°C, the specific heat from Table 2.2 would suggest that the local SAR for human fat is 1.25 W/kg.

These values calculated from the table are referred to as local SAR quantities. The local SAR measures the power absorbed by usually 1 g or 10 g of the material in consideration.
When the entire human body is considered or any other mass with a distributed specific heat, the averaged SAR is used for measurements. This will provide more accurate results when the entire body is absorbing the EM radiation.

Tissue phantoms are often used for measurements in determining the safety limitations of exposure times and power levels. These phantoms are created to closely resemble the SAR of actual tissue and are often used in experiments regarding cellular telephones and base stations. These physical phantoms achieve the desirable conductivity and relative permittivity by mixing together different materials (water, Sodium Chloride, etc.) [10]. Simulation software can perform a similar task by synthesizing materials within the model to also match the desirable conductivity and relative permittivity.

### 2.1.3. DIELECTRIC AND INDUCTION HEATING THERMOTHERAPY

Heat can be applied to biological tissue with the intention of medical treatment. Heated tissue can relax muscles and joints and assist in the body’s recovery time. Excessive heat, however, can be detrimental to living cells. Most living cells will die starting at 42.5°C [11]. Water is a large percentage of healthy cells, allowing for microwave heating applicators to efficiently heat these tissues. When heat is applied in a controlled manor, diseased tissue could be eliminated by taking advantage of the fact that tumors and other

<table>
<thead>
<tr>
<th>Tissue</th>
<th>Specific Heat Capacity (J kg⁻¹ °C⁻¹)</th>
<th>Density (kg m⁻³)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Skeletal muscle</td>
<td>3470</td>
<td>70</td>
</tr>
<tr>
<td>Fat</td>
<td>2260</td>
<td>940</td>
</tr>
<tr>
<td>Bone, cortical</td>
<td>1260</td>
<td>1790</td>
</tr>
<tr>
<td>Bone, spongy</td>
<td>2970</td>
<td>1250</td>
</tr>
<tr>
<td>Blood</td>
<td>3890</td>
<td>1060</td>
</tr>
</tbody>
</table>

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Tissue phantoms are often used for measurements in determining the safety limitations of exposure times and power levels. These phantoms are created to closely resemble the SAR of actual tissue and are often used in experiments regarding cellular telephones and base stations. These physical phantoms achieve the desirable conductivity and relative permittivity by mixing together different materials (water, Sodium Chloride, etc.) [10]. Simulation software can perform a similar task by synthesizing materials within the model to also match the desirable conductivity and relative permittivity.

### 2.1.3. DIELECTRIC AND INDUCTION HEATING THERMOTHERAPY

Heat can be applied to biological tissue with the intention of medical treatment. Heated tissue can relax muscles and joints and assist in the body’s recovery time. Excessive heat, however, can be detrimental to living cells. Most living cells will die starting at 42.5°C [11]. Water is a large percentage of healthy cells, allowing for microwave heating applicators to efficiently heat these tissues. When heat is applied in a controlled manor, diseased tissue could be eliminated by taking advantage of the fact that tumors and other

<table>
<thead>
<tr>
<th>Tissue</th>
<th>Specific Heat Capacity (J kg⁻¹ °C⁻¹)</th>
<th>Density (kg m⁻³)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Skeletal muscle</td>
<td>3470</td>
<td>70</td>
</tr>
<tr>
<td>Fat</td>
<td>2260</td>
<td>940</td>
</tr>
<tr>
<td>Bone, cortical</td>
<td>1260</td>
<td>1790</td>
</tr>
<tr>
<td>Bone, spongy</td>
<td>2970</td>
<td>1250</td>
</tr>
<tr>
<td>Blood</td>
<td>3890</td>
<td>1060</td>
</tr>
</tbody>
</table>

When the entire human body is considered or any other mass with a distributed specific heat, the averaged SAR is used for measurements. This will provide more accurate results when the entire body is absorbing the EM radiation.

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<td>3890</td>
<td>1060</td>
</tr>
</tbody>
</table>
malicious cells have a higher concentration of water, thus heating faster [11]. The medical community refers to devices that supply the heat as applicators. These antennas can utilize two different types of heating principles: dielectric heating or inductive heating. Both principles rely on the movement created by subatomic particles; the difference lies in how the movement occurs.

Dielectric heating is usually considered for materials that are not highly conductive. A constantly changing polarization is created inside of the material’s particles when an alternating electric field is applied. Heat energy is given off as a result of friction in the material when the particles move on a molecular scale. Non-polarized particles will have shifting electrons creating this friction. Polarized molecules, such as water, have an even greater heating effect because the molecule is already polarized and will rotate when subjected to the alternating field.

Figure 2.3. Parallel plate capacitor
If a dielectric is subjected to an applied voltage, such as the parallel plate capacitor shown in Figure 2.3, the power loss in the material can be expressed in watts, as shown in (2.7). This power loss directly correlates to the heat generated in the material. In this relationship, $R_{ESR}$ is the equivalent series resistance of the material inside of the capacitor.

$$P_d = \frac{V_{\text{rms}}^2}{R_{ESR}}$$  \hspace{1cm} (2.7)

Figure 2.4 shows a typical control diagram for a capacitive heating device. This device can generate 1.5 kW of maximum output power and will operate at 8 MHz [12]. Because the goal is usually to heat an area of tissue or a tumor deep inside the body beyond the skin, the skin must be kept cool with water during the process. The skin effect states that the human skin will absorb most of the energy, thus it must be cooled to allow for deeper heating from the exposure time.

Just as dielectric heating relies on the properties of relative permittivity, inductive heating is dependent upon a material’s conductivity and relative permeability. When a

![Figure 2.4. An 8 MHz RF capacitive heating device. Extracted from [12]]
Figure 2.5. Toroidal ferrite core with two poles with embedded metallic sphere in tissue phantom. Extracted from [13]

Conductive material is subjected to an alternating magnetic field, such as a coil of wire carrying an alternating current wound around a ferrite core, the conductive material will generate eddy currents to oppose the field, as stated in Lenz’s law. These internal currents encounter some resistance in the conductor, resulting in power loss and Joule heating.

Figure 2.5 demonstrates this property through the use of two wound ferrite cores on opposite sides of a phantom tissue with an implanted metallic sphere. The alternating magnetic field applied to the material has the largest influence on the sphere. The sphere experiences power loss from the eddy currents and heats itself and the surrounding tissue.

The configuration demonstrated in Figure 2.5 can be used to create a comparison between dielectric and induction heating applicators. The results of these application types are shown as thermal contours in both Figure 2.6 and Figure 2.7. Both applicator experiments were performed with similar input power levels, phantom tissues, and a metallic sphere implanted in the material.
Figure 2.6. Thermographic view of experimental results with capacitive applicator. Extracted from [14]

Figure 2.7. Thermographic view of experimental results with inductive applicator. Extracted from [14]
It is important to note the presence of the fat layer in this phantom material. Because the capacitive applicator uses an electrode pair similar to a parallel plate capacitor, the thermal effects are distributed among the entire exposed volume, including this fat layer. The capacitive technique does not allow for isolation of certain areas and is effective at heating large masses. The fat layer can be seen around the Agar phantom and the metallic sphere in Figure 2.6.

With the inductive heating technique, only the metallic sphere is heated, thus conducting heat to the surrounding tissue. There is not an increased temperature measurement in the fat layer in Figure 2.7. This inductive heating technique is useful for isolating tissues that require thermotherapy while not effecting the nearby tissues [14].

2.1.4. RF Hyperthermia and RF Ablation

RF hyperthermia techniques can be used to destroy malignant tissues while causing minimal harm to surrounding healthy tissues. Hyperthermia is a process of overheating a tissue to invoke a desired process. In the case of biological tissue, this operation is performed to kill unhealthy cells and tumors. The desired temperature patterns can be achieved by selecting the correct type of RF applicator, capacitive or inductive, depending on the damaged area. Inductive heating techniques find more usage because of the ability to isolate tissues, however dielectric heating should not always be discounted.

Tumors are generally more susceptible to hyperthermia because of their growth rate. Because of this growth rate, tumors have a greater water density, allowing greater heat transfer than normal tissues. Tumor growth also outpaces blood vessel development, not allowing quick heat displacement through the bloodstream. These are also a hypoxic region due to the lack of blood and oxygen delivery, creating more hyperthermia sensitivity [11].
While hyperthermia can be used as a general term to describe the process of heating a tissue, ablation usually refers to the removal of surface material. RF ablation performs this task with an interstitial device that utilizes the skin effect to remove surface material from a tissue. When the internal device radiates power, it heats the tissue in contact according to the radiation pattern.

The most common medical operation with RF ablation is performed in the heart [15]. Cardiac ablation is a procedure to remove a layer or layers of tissue that could be causing arrhythmias. RF ablation has also been used in other parts of the body to open pathways for better fluid flows: blood, urine, and air paths being the most common.

A majority of ablation and hyperthermia applicators operate in frequency bands around 900 MHz and 2 GHz because the FCC has allocated these bands for scientific, medical, and industrial usages. With the availability of these bands, high-powered generators have continued to improve in these areas, further allowing the medical community to develop applicators of the same relative size. Regardless of the frequency, interstitial ablation devices usually create an ablation zone around the device no more than a quarter wavelength away [15]. Figure 2.8 shows the ablation zone of a 1.9 GHz device.

*In vivo* simulations showed that after supplying 42 W for 10 minutes, the area encompassing the -15 dB zone was heated to 60ºC. This temperature distribution can be effective for killing malicious cells within this radius.
Figure 2.8. Zones showing power loss of 1.9 GHz ablation catheter inside of tissue. Extracted from [15]

Figure 2.9. Comparison of -25 dB loss contours from 1.9 GHz and 10 GHz ablation systems. Extracted from [15]
Similar ablation zones can be accomplished by also using higher frequency devices. Although higher frequencies will penetrate less into tissues, it has been shown experimentally that heat conduction plays a more important role at these frequencies [15]. The lesser skin depth produces higher temperatures at the tissue-applicator interface, but this creates a higher heat transfer rate. The -25 dB power zones for two different frequencies are shown in Figure 2.9. It is known that the effective heating zone is within \( \lambda/4 \) as reflected in this figure.

Experimental results have shown only a 9% difference in ablation areas of 1.9 GHz and 10 GHz systems with 42 W input for 10 minutes [15]. These ablation zones are shown in Figure 2.10. It is important to note the more intense scarring around the device for the 10 GHz system. This is attributed to the lower penetration depth, resulting in more damaged tissue closer to the device.

The tissue-applicator interface is reported to reach 100°C in 45 sec in the 10 GHz system [15]. This magnitude is not a major concern because this interface should be located within the tumor ablation region. It is this high temperature that creates the resulting large area. The high temperature creates a larger heat gradient, allowing more heat to travel towards the outer region. It may appear that the 1.9 GHz system performs better at heating the entire area, but the experimental results disprove this idea.
Figure 2.10. Photographs of the ablation zone obtained using 42 W of microwave power at (a) 10.0 GHz for 5 min, (b) 1.9 GHz for 5 min, (c) 10.0 GHz for 10 min, and (d) 1.9 GHz for 10 min. Extracted from [15]

2.2. RF HEATING OF INORGANIC MATERIALS

The fundamental difference between heating organic and inorganic materials is the presence of water molecules. While not all living tissues has high concentrations of water, the average water content is enough to generate heat from the rotating molecular dipole in the alternating electric field. The previous section’s examples mainly consisted of dielectric heating because of these molecules, however inorganic materials do not always have this trait. While some materials, such as common materials like PVC and nylon still utilize
dielectric heating because of their molecular dipoles, most inorganic substance that require heating are metals [16].

2.2.1. DIELECTRIC HEATING

RF welding is the process of joining plastics together by exposure to RF fields. This process can only be used on plastic molecules that contain a molecular dipole [17]. The most commonly used materials for RF welding are as follows: PVC, nylons, polyurethane, polyethylene terephthalate (PET), and polyvinylidene chloride (PVDC) [16], [17], [18].

Figure 2.11 shows a typical welding process of two work pieces. The electrodes act as a capacitor with the work pieces placed between. The alternating electric field between the two electrodes causes the molecular dipole in the material to rotate, causing friction and heat. All work pieces have a desired heating point past their melting temperature that must be achieved in order to create a clean weld. It is undesired to have a partial weld where the two materials do not fuse at the molecular level. Although Figure 2.11 shows a dividing line in the product after the weld, the true result is on uniform material, as if it were never separated.

![Figure 2.11. Schematic representation of RF welding (before the application of RF energy on the left and afterwards on the right). Extracted from [18]](image-url)
2.2.2. **Induction Heating**

Induction heating principles were first observed with Faraday. Faraday discovered induced currents by an alternating magnetic field which Maxwell later developed into the unified theory of electromagnetism. Joule was able to describe the heat produced by a current in a conductor, all together establishing the fundamental principles of induction heating [19]. Induction heating continues to grow in industrial and medical applications because of the efficiency and clean process [20], [21]. Typical industrial applications operate at line frequency up to 5 kW while medical applications vary in the megahertz range and typically use less than 2 kW [22].

Heat is mainly created by eddy currents and magnetic hysteresis [23]. The eddy currents induced by the alternating magnetic field encounter some surface resistance, mainly within the skin depth of the work piece. These currents and resistance directly correlate with the $I^2R$ losses in the material and create Joule heating on the surface. Because of the thermal conductivity, this heat spreads locally and the volume of the heated piece can be controlled by input power, coil sizing, and exposure times. Figure 2.12 is extracted from the referenced text and illustrates a simple configuration of a heating coil, work piece, and the eddy currents on the material’s surface.

![Figure 2.12](image)

*Figure 2.12 Typical arrangement of an induction heating system in a longitudinal flux configuration showing general view and top view. Extracted from [19]*
2.3. FREQUENCY SELECTIVE SURFACES

Frequency selective surfaces (FSS) are typically formed by a two-dimensional array of metallic patterns printed on a dielectric substrate [24]. Although a FSS does not have to be made from a good conductor, it is usually fabricated with copper and simulated with perfect electrical conductors when dealing with reflective antennas [25]. Also, it does not necessarily have to be a two-dimensional structure. A simple example of a three-dimensional material can be made by interweaving separate 2-D planes orthogonally to each other [26].

The functionality of a FSS is to absorb, reflect, or pass electromagnetic waves of various frequencies. When fashioned in specific ways, the FSS can absorb EM radiation of certain frequencies like a lossy material, while also reflecting and passing other frequencies when incident to the material. These three modes depend on how the EM radiation induces currents on the surface of the material. When the surface is absorbing EM radiation, currents are created on the material and the absorbed wave is lost as heat. Reflection happens at wavelengths that are too large to induce a significant amount of current on the material. Depending on the area of the material and empty space, transmission can happen at smaller wavelengths [26]. These wavelengths are of course relative to the dimensions of the fabricated FSS.
Figure 2.13 Mesh style frequency selective surface. Extracted from [27]

Figure 2.14. Patch style frequency selective surface. Extracted from [27]
2.3.1. **Classical Example**

The two basic types of FSS are meshes and patches [27]. This classical example will outline the functionality and theory behind both structures, with all material in this discussion originating from Munk [27].

Mesh and patch FSS are complimentary structures that demonstrate Babinet’s principle. The complimentary structure can be seen graphically in Figure 2.13 and Figure 2.14. Babinet’s principle states that these two structures’ frequency responses will be exactly inverse, shown in Figure 2.15. Figure 2.13 shows a mesh type of FSS and will exhibit band-pass characteristics. If a frequency sweep were performed on the device, a response like Figure 2.15 (a) is expected. At a design frequency of $f_o$, all the incident signal will be transmitted through and rejection will happen in the remainder of the spectrum. The patch style in Figure 2.14 however operates as a band-stop filter. All frequencies in the spectrum will be passed except for the design frequency, shown in Figure 2.15 (b).

For both surfaces, the lattice constant $a$ and the width/spacing term $w$ are the main variables when designing a frequency response. These two design variables influence the inductance and capacitance that the surface will exhibit. Each surface can be represented as an equivalent circuit acting as a shunt impedance along a transmission line. For the band-pass mesh, this impedance is the parallel combination of an inductor and a capacitor, seen in Figure 2.16. The band-stop patch model is a series combination of an inductor and a capacitor, seen in Figure 2.17.
The parallel LC combination for the band-pass mesh will have the following shunt impedance and admittance:

\[
Z_{\text{shunt}} = \frac{1}{\frac{1}{j\omega L} + j\omega C}
\]

\[
Z_{\text{shunt}} = \frac{j\omega L}{1 - \omega^2 LC}
\]

\[
Y_{\text{shunt}} = \frac{1 - \omega^2 LC}{j\omega L}
\]
When the desired frequency nulls the reactance of both components, there will be zero admittance along the shunt path. This is like an open circuit with high impedance that allows all transmission down the transmission line, reflecting no energy.

The series LC combination for the band-stop patch will have the following shunt impedance:

\[
Z_{\text{shunt}} = j\omega L + \frac{1}{j\omega C}
\]

\[
Z_{\text{shunt}} = \frac{1 - \omega^2 LC}{j\omega C}
\]
When the desired frequency nulls the reactance of both components, there will be zero impedance along the shunt path. This is like a short circuit with low impedance that will reflect all energy back towards the input and not allow any transmission along the line.

2.3.2. COMMON APPLICATIONS

Frequency selective surfaces first found common use in RADAR stealth. In 1952 Winfield Salisbury filed a patent for a device that could effectively absorb EM radiation at certain frequencies [28]. Named the Salisbury screen, it could be placed on vehicles prone to being found via RADAR, mainly airplanes. An exploded view of a Salisbury screen is shown in Figure 2.18.
The screen is a dielectric material placed against a conducting material with the dielectric’s open face exposed to the incident EM radiation. Every wave that interacts with the dielectric will have a reflected component, but when the dielectric thickness is \(\lambda/4\) of the incident wave, there is a perfect destructive wave reflected from the conductor. When the portion of the incident wave that is not reflected from the dielectric travels through the screen it is reflected at the conductor and then passes back through the dielectric. This component of the wave is now 180° out of phase with the original waveform, completely canceling the total reflection [28].

When used for stealth, the Salisbury screen is only effective at one frequency. Should the scanning device shift to a new carrier frequency, that screen would be useless. Outside of stealth, this technique is still useful for applications in antenna measurements and antenna array design.
The microwave oven is a typical household device that heats food by dielectric heating. This type of heating is obtained by exposing a dielectric material to a varying electric field. In the case of the microwave oven, the water molecules in food are forced to rotate in the alternating field, creating microscopic friction. This friction generates the heat inside of foods. Water molecules have an electric dipole due to the bonding angle of Hydrogen and Oxygen. This dipole is what causes the molecule to rotate in the changing field. The microwave oven operates at 2.45 GHz which can easily pass through the glass viewing door. Should this happen, it could cause the same effect to water molecules outside of the walls of the microwave oven. Five of the microwave oven walls are continuous metal, which do not allow the EM radiation to escape. In order to still see the contents in the oven, a FSS is placed on the glass door. This FSS is designed to reflect the operating frequency of the device while still allowing shorter light wavelengths to pass. Typical ovens have the mesh style surface with holes at 1 mm, 120 times smaller than the wavelength of operation [28]. The visible light wavelengths are approximately 2,000 times smaller than these holes and are able to pass.

2.3.3. FSS FOR MICROWAVE POWER TRANSMISSION

An interesting publication by Z. H. Wang is referenced in [25]. This journal entry describes how a frequency selective surface was utilized to eliminate the second and third harmonics of a rectenna system. With significantly high power, these undesirable harmonics can radiate and interfere with nearby equipment and components.

A FSS was designed to eliminate the second harmonic at 4.9 GHz and the third harmonic at 7.35 GHz of a microwave power transmission system. The dimensions for the FSS were calculated using the methodologies described in [30],[31],[32]. For the proposed
gridded-square array in Figure 2.19 (a), the dimensions are unique for each frequency range. The gridded-square array can only accomplish resonance at one frequency, but multiple layers of each design can be implemented to achieve the desired result. For 4.9 GHz: $p = 27.0$ mm, $w_1 = 0.2$ mm, $w_2 = 3.5$ mm, $d = 21.8$ mm. For 7.35 GHz: $p = 24.8$ mm, $w_1 = 0.2$ mm, $w_2 = 7.0$ mm, $d = 22.0$ mm. The double-square array is illustrated in Figure 2.19 (b) and has two resonant frequencies. This functionality would allow the single structure to eliminate both harmonics without the need of additional space or complexity. To achieve high attenuation at 4.9 GHz and 7.35 GHz: $p = 30.6$ mm, $w_1 = 0.2$ mm, $w_2 = 0.2$ mm, $d_1 = 19.0$ mm, and $d_2 = 11.9$ mm.

Figure 2.19. FSS and equivalent circuit models for (a) gridded-square array and (b) double square array. Extracted from [25]
These FSS designs were fabricated and measured in an anechoic chamber. Because of resources, the gridded-square array was only designed for the second harmonic at 4.9 GHz while the single implementation of the double-square array is able to attenuate the second and third harmonics naturally by geometry. Both surfaces were tested at multiple incident angles to determine the effectiveness of the surface and its placement in the system.

The gridded-square array was able to achieve 20 dB attenuation at the 4.9 GHz harmonic. The double-square array obtained 10 dB attenuation at each of the harmonics. The insertion loss at 2.45 GHz for both surfaces is less than 0.5 dB. The paper cites both of these solutions valid in practice, depending on the most important harmonics.

Figure 2.20. Frequency sweep of the system utilizing the gridded-square array. Extracted from [25]
2.3.4. CORRUANTED SPLIT RING RESONATOR

The following discussion is a summary of K. M. Z. Shams findings regarding a wider band split ring resonator (SRR) [33]. The bulk of this discussion is extracted from Chapter 5 of his Ph.D. dissertation.

This research deals with a proposed corrugated split ring resonator to function as a stopband filter with more bandwidth than the conventional SRR. These two configurations are illustrated in Figure 2.22 (a) and (b).
Figure 2.22. Labeled dimensions on (a) conventional SRR and (b) corrugated SRR.
Extracted from [33]
Both structures were simulated in a waveguide environment on similar dielectrics of comparable size, the largest dimension being $L_1 = 10$ mm. The transmission and reflection characteristics of both devices were recorded and are shown in Figure 2.23.

![Figure 2.23. S-parameters for conventional and proposed SRR. Extracted from [33]](image)

For the conventional resonator, $S_{21} \approx 0$ dB at most of the sweep with very low transmission around 2.15 GHz. This idea is also represented in the conventional resonator’s $S_{11}$ plot, with significant reflection at 2.15 GHz. This suggest that the surface is acting as a band stop filter around 2.15 GHz. The corrugated resonator has $S_{21} \approx 0$ dB for most of the sweep and reaches near -40 dB around 2.45 GHz. As before, $S_{11}$ is nearly the inverse, suggest reflection around 2.45 GHz. This resonator also acts as a band stop filter, only at a higher frequency for the same footprint size. The document [33] emphasizes that the bandwidth of the conventional resonator is 4.7% while the new corrugated resonator is
7.3%. These bandwidth calculations are based on the widths of the band below -10 dB for each respective transmission term.

The text further experiments with an equivalent circuit model for the new resonator and simulates different design parameters to validate the model. Figure 2.24 introduces a new term \( l \) to the design variables for the SRR. This variable is used to represent the effective length of the resonator. The equivalent circuit model in Figure 2.25 can be used to simulate and tune the SRR once a correlation is made between the geometry and the circuit model.

Figure 2.24. Corrugated SRR showing effective length of the structure and design variables. Extracted from [33]

Figure 2.25. Equivalent circuit model for the corrugated SRR. Extracted from [33]
The corrugated SRR was perturbed by one variable over several trials to experimentally determine how each design choice influenced the frequency response of the surface. While operating under item 1 in Table 2.3, Figure 2.26 was created. This figure shows the magnetic and electric fields on the resonator at 2.5 GHz. It is clear that the ring half of the structure acts as the inductance because of the high magnetic field along the effective length path. The corrugated teeth behave as a capacitance, exhibited by the strong electric field intensity near the gaps.

In Table 2.3, $\varepsilon_r$ is the relative permeability of the substrate beneath the SRR, $s$ is the width of the trace in mm, $g$ is the width of the gaps in mm, $n$ is the number of teeth pairs, $A$ describes how the SRR is arranged in an array, $d$ is the periodicity of the SRR in the array, and $l$ is the effective length path along the SRR.

Item 1 is used as a baseline. When the number of capacitive teeth decrease, the frequency slightly increases. With more capacitive teeth, the frequency slightly decreases. With a smaller trace, there is a more significant increase in the frequency while a wider trace has a more significant decrease in the frequency. The spacing of the gaps follows the same trend as the trace width. If the periodicity of the grid decreases, the operating frequency increases.

An interesting trend that the paper mentions is that as the frequency increases, the deviation between the circuit model and the HFSS simulation decreases. The author mentions that this is probably due to the same bandwidth remains, but the fractional calculation reduces because of the higher frequency.
Figure 2.26. Magnetic and electric field strengths present on the surface of the corrugated SRR at 2.5 GHz. Extracted from [33]
Table 2.3 Trial Variations of the Corrugated Split Ring Resonator and Frequency Band, Extracted from [33]

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<tr>
<th>Item</th>
<th>$\varepsilon_r$</th>
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<th>g mm</th>
<th>n</th>
<th>$A$ mm$^2$</th>
<th>$d$ mm</th>
<th>$l$ mm</th>
<th>$f_L$ GHz</th>
<th>$f_H$ GHz</th>
<th>% deviation</th>
<th>$f_L$ GHz</th>
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<td>3</td>
<td>4.7x4.7</td>
<td>11</td>
<td>14.3</td>
<td>5.95</td>
<td>5.98</td>
<td>0.5</td>
<td>6.66</td>
<td>6.51</td>
<td>2.3</td>
</tr>
</tbody>
</table>
CHAPTER 3. ALLOY CHARACTERIZATION THROUGH TRANSMISSION LINE SIMULATION AND EXPERIMENTATION

The RF wireless power absorbed by the Lanthanum-Nickel alloy can be calculated in the form of $I^2R$ losses. The surface resistance, $R_s$, must be determined, as it is substituted for $R$ in the power calculation. The surface resistance, $R_s$, for a particular case depends on the frequency of the RF source, $f$, the material’s electrical conductivity, $\sigma$, and relative permeability, $\mu_r$. The values of the latter two terms for the alloy are unknown. In order to continue, a methodology was developed to isolate the conductivity and relative permeability of the alloy. This process revolves around fabricating a transmission line with the material. An HFSS model of a Copper transmission line on FR4 is used to isolate the loss component due to the dielectric. With a mathematical model for the FR4 dielectric loss, actual lines were fabricated from the alloy and the S-parameters were measured. This loss is then removed from the experimental data, which is brought back to HFSS for curve-fitting via a parametric sweep of the unknown conductivity and relative permeability.

![Flow chart diagram of determining conductivity and relative permeability](image)

Figure 3.1 Flow chart diagram of determining conductivity and relative permeability
3.1. TRANSMISSION LINE DEVELOPMENT AND SIMULATION

To estimate the above parameter values, a transmission line type setting of simulation and measurement was considered for its simplicity. Since a microstrip transmission line can be easily modeled and simulated in HFSS, such structures were considered for both simulation as well as experiments. Given the material under consideration was granular a narrow long channel was created for the experimental case where the material grains could be placed and consolidated as much as possible.

For the experimental specimen copper clad single sided 1.6 mm thick FR4 PCBs (Printed Circuit boards) were considered. Each board was cut into a 30 mm x 30 mm square where a narrow long channel was milled that ran along the length of the square in the center of the surface. The groove depth was 0.3 mm. A machining schematic of the described board is shown in Figure 3.2.

On the simulation side, the loss introduced from the FR4 substrate was determined through HFSS simulation by modeling the fabricated board and filling the groove volume with Copper. Two simulations were performed to determine the FR4 loss: one model with the FR4 native material definition, loss tangent of 0.02, and a second model with a modified FR4 definition, loss tangent of zero. Figure 3.3 shows the HFSS model used to arrive at the set of S-parameters. The model was created from the dimensions in Figure 3.2.

S-parameters were simulated for both Copper transmission line models from 0.3 – 6.0 GHz, with port impedances at 50 Ω. With these two sets of S-parameters, the mismatch
Figure 3.2 Machining schematic for the two-port FR4 board.
Figure 3.3 HFSS model of the fabricated FR4 boards

and trace loss can be determined and used to isolate the loss experienced in the
dielectric. The S-parameters obtained from the substrate containing the loss tangent have a
mismatch loss as well as losses incurred by the trace and dielectric. The second simulation
only contains a mismatch loss and trace loss. These results are plotted in Figure 3.4 and
Figure 3.5. S_{11} approaches a minimum in both boards at 2.765 GHz and 5.400 GHz. From
the model dimensions, the effective permittivity from (3.1) is \( \varepsilon_{\text{eff}} = 3.32 \), yielding guided
wavelengths of 59.51 mm and 30.47 mm respectively. It is then implied that at these two
frequencies there will be a high return loss due to the trace length being 30 mm. The same
will be true for any frequency whose guided wavelength is a multiple of the board length.

\[
\varepsilon_{\text{eff}} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2\sqrt{1 + 12d/W}}
\]  
(3.1)
Figure 3.4 Simulated $S_{11}$ from Copper trace on FR4 board

Figure 3.5 Simulated $S_{21}$ from Copper trace on FR4 board
For the remainder of the discussion, all terms containing an ‘x’ in the subscript denote origin from the model without a loss tangent and S-parameter terms labeled with a lowercase ‘s’ are numerical whereas an uppercase ‘S’ is in decibels.

MATLAB was used to determine the mismatch loss exhibited in $S_{21x}$ through (3.2). Because $S_{21x}$ only contains the mismatch loss and trace loss, (3.3) is applied to isolate the trace loss.

$$\text{Mismatch Loss}_x = 10 \log_{10}(1 - |s_{11x}|^2) + 10 \log_{10}(1 - |s_{22x}|^2) \quad (3.2)$$

$$\text{Trace Loss} = S_{21x} - \text{Mismatch Loss}_x \quad (3.3)$$

With the mismatch loss and trace loss both determined, the dielectric loss is isolated by removing the two mentioned values from $S_{21}$. This calculation is described in (3.4).

$$\text{FR4Loss} = S_{21} - \text{Trace Loss} - \text{Mismatch Loss} \quad (3.4)$$

The dielectric loss from the FR4 board can now be removed from any $S_{21}$ data set between 0.3 – 6.0 GHz obtained from this milled board. The dielectric loss is plotted in Figure 3.6 along with the mismatch loss and trace loss determined over this frequency range. All three of these terms create $S_{21}$, also shown in the figure, as a rearrangement of (3.4). A linear model of the dielectric loss was derived using MATLAB’s least square fitting with an $R^2$ value of 0.9968. This linear approximation is described in (3.5). The loss model and linear approximation are shown in Figure 3.7 for comparison.

$$\text{FLLM}(f) = -0.1037f + 0.01845 \quad (3.5)$$
Figure 3.6 Individual components of $S_{21}$ in Copper trace on FR4 board

Figure 3.7 Isolated FR4 loss and linear fit to the loss model
The FR4 Linear Loss Model (FLLM) was then applied to the $S_{21}$ of the model containing the loss tangent for validation. This is done by subtracting the FLLM from the simulated trace. As seen in Figure 3.8, when the FLLM is applied to $S_{21}$, it approaches $S_{21x}$, the simulation without the loss tangent. This is the desired result for validation.

The maximum difference between $S_{21x}$ and $S_{21}$ (FLLM) over the 20:1 frequency range is 0.02 dB and occurs at 0.3 GHz. This minute difference, along with the high $R^2$ value from the fitting, demonstrates how well the technique performs.

3.2. ANALYSIS OF EXPERIMENTAL RESULTS

The fabricated board was subject to fill the groove with the alloy material and take $S$-parameter measurements. The material is granular after synthesis and was pressed into the groove by a compressive cylinder. SMA connectors were attached after the material
was pressed into the groove allowing the S-parameters to be measured on the 2-port network. Many fabricated boards were created, however of all the boards machined, loaded, and measured, only the three most confident samples were used during this analysis. Some measurements showed major disagreements which were ultimately contributed to the lack of contact between the granular material in the grooves along with how firmly the SMA pins made contact with the material in the trace. Figure 3.9 shows the three selected samples. The S-parameters from the three selected samples first had their respective mismatch removed. The mean of the resulting transfer function terms is then considered to be the experimental measurement of $S_{21}$ for the alloy trace. The FLLM is then applied to this trace, leaving only the trace loss due to the alloy material. The resulting traces are shown in Figure 3.10.

![Figure 3.9 Measured S-parameters from alloy trace on FR4 board. Courtesy of Michael Brown, Savannah River National Laboratory](image-url)

45
This loss described in Figure 3.10 is derived from a modified version of (3.4) where the FR4 loss is now defined as the FLLM. This rearrangement is shown in the equation below.

\[
\text{Trace Loss} = \frac{1}{3} \sum_{i=1}^{3} (S_{21,i} - \text{Mismatch Loss}_i) - \text{FLLM}(f) \quad (3.6)
\]

3.3. DETECTION OF CONDUCTIVITY AND RELATIVE PERMEABILITY

The alloy trace loss in the experimental transmission lines can be mimicked with an HFSS model to determine the conductivity and relative permeability of the material. The model shown in Figure 3.3 is reused with an artificial conductor in the groove whose conductivity and relative permeability are defined as variables in the model. HFSS allows
the conductor to have a surface roughness when it is defined as a finite conductivity boundary, which was set to 16 µm. This value was selected because the granular alloy material was sieved through a 16 µm geometry before being pressed into the transmission line grooves. This value is assumed to be the nominal surface roughness if the largest grains provide the largest source of roughness. This idea is conveyed in Figure 3.11 with a cross-sectional view of the groove on the center of the transmission line boards. If the groove is pressed with the granular material, the most surface roughness would be exhibited from the largest granular pieces near the surface of the trace.

![Cross-sectional side view of groove to illustrate surface roughness from largest alloy grains](image)

Figure 3.11 Cross-sectional side view of groove to illustrate surface roughness from largest alloy grains

The HFSS optimization tool was used to retrofit the experimental data without the FLLM in Figure 3.10. The conductivity and relative permeability were assigned to variables that the optimization tool could tune to achieve the experimental trace loss. This technique was performed on the response without the applied FLLM because HFSS could quickly plot the results for removing the mismatch, whereas applying the FLLM would require post-processing in MATLAB. This is justified in that by applying the FLLM, it is a removal of a linear function, providing the same result before or after subtracting the model.
The loss seen in Figure 3.10 is considered undesirably non-linear after 3.0 GHz. For better retrofitting, the optimization tool was given two points to fit: at 0.3 GHz and 3.0 GHz. Ignoring the trace after 3.0 GHz will limit the material characterization to this frequency range; all future testing will be within this range. These goals are extracted from the trace without the FLLM in Figure 3.10.

| Frequency | $S_{21} - 10 \log_{10}(1 - |s_{11}|^2) - 10 \log_{10}(1 - |s_{22}|^2)$ |
|-----------|--------------------------------------------------------------------------|
| 0.3 GHz   | -0.3576 dB                                                               |
| 3.0 GHz   | -1.6450 dB                                                               |

The final optimized response from HFSS, given the goals in Table 3.1, is shown in Figure 3.12. The optimization process adjusted the conductivity and relative permeability to achieve this result. Figure 3.12 shows the comparison between the experimental data and the HFSS model. The software chose the material parameters listed in Table 3.2. The retrofitting process matches well at the edges of the frequency range, as designed, but has some error in the center of this band. The largest disagreement is 0.046 dB and occurs at 1.21 GHz. This minimal error demonstrates that the material parameters are closely approximated by the values in Table 3.2.

<table>
<thead>
<tr>
<th>Surface Roughness – $\delta_{sr}$</th>
<th>Conductivity – $\sigma$</th>
<th>Relative Permeability – $\mu_r$</th>
</tr>
</thead>
<tbody>
<tr>
<td>16 $\mu$m</td>
<td>$1 \times 10^6 \ S/m$</td>
<td>110</td>
</tr>
<tr>
<td>0 $\mu$m</td>
<td>$2.5 \times 10^5 \ S/m$</td>
<td>110</td>
</tr>
</tbody>
</table>
Figure 3.12 Comparison of experimental trace loss and optimized parameter retrofitting before and after applying FLLM

Table 3.2 also contains a set of pseudo-parameters if the surface roughness is ignored. These values were determined with the same optimization scheme when the surface roughness was set to zero. These values are not considered to be the actual parameter values for the material, however all future simulations utilize these terms because of simulation speed.

These values for conductivity and relative permeability are not unique, however. The ratio $\mu_r/\sigma = 110 \times 10^{-6}$ can be used to determine any combination of the two parameters that yields the same plot. If the surface roughness is ignored, $\mu_r/\sigma = 440 \times 10^{-6}$. The values listed in Table 3.2 will be considered true for the remainder of this work, with all remaining discussion ignoring the surface roughness for simulation speed and convenience.
CHAPTER 4. SIMULATED ENERGY TRANSFER WITH LOG-PERIODIC DIPOLE ARRAY

A log-periodic dipole array was selected as the first antenna to transfer energy to the material. A HyperLOG® 7060 LPDA was initially purchased for preliminary testing with the alloy material. This selection was primarily based on the directivity and gain of a LPDA, but also for its wide bandwidth during the testing stage. In order to determine an expected result from this antenna, a LPDA with similar specifications was first modeled in HFSS using documented parameters from the datasheet of the array. An alloy material slab is then introduced in front of the LPDA model and the results were recorded. By exporting the current density induced on the material slab, a post-processing technique developed in MATLAB can be used to calculate the rise in temperature of the material.

4.1. LOG-PERIODIC DIPOLE ARRAY DESIGN

The HyperLOG® 7060 LPDA specifications are shown in Table 3.1. A design from the Carrel curves was followed to achieve these design parameters [34]. The design variables $\tau$ and $\sigma$ were chosen to be 0.822 and 0.149 respectively. These values were used to determine the dimensions with the longest element determined by the lowest frequency.

\[
\tau = \frac{L_n}{L_{n-1}}
\]  

\[
\sigma = \frac{D_{n-1}}{2L_{n-1}}
\]
Table 4.1 HyperLOG® 7060 LPDA Specifications

<table>
<thead>
<tr>
<th>Specification</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range</td>
<td>700 MHz – 6 GHz</td>
</tr>
<tr>
<td>Max. Transmission Power</td>
<td>100 W (400 MHz)</td>
</tr>
<tr>
<td>Nominal Impedance</td>
<td>50 Ω</td>
</tr>
<tr>
<td>VSWR (typ.)</td>
<td>&lt; 1:2</td>
</tr>
<tr>
<td>Gain (typ.)</td>
<td>5 dBi</td>
</tr>
</tbody>
</table>

Table 4.2 Log Periodic Dipole Array Dimensions (mm)

<table>
<thead>
<tr>
<th>#</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>8</th>
<th>9</th>
<th>10</th>
<th>11</th>
<th>12</th>
</tr>
</thead>
<tbody>
<tr>
<td>L</td>
<td>202.00</td>
<td>166.04</td>
<td>136.49</td>
<td>112.19</td>
<td>92.22</td>
<td>75.81</td>
<td>62.31</td>
<td>51.22</td>
<td>42.10</td>
<td>34.61</td>
<td>28.45</td>
<td>23.39</td>
</tr>
<tr>
<td>D</td>
<td>60.20</td>
<td>49.48</td>
<td>40.67</td>
<td>33.43</td>
<td>27.48</td>
<td>22.59</td>
<td>18.57</td>
<td>15.26</td>
<td>12.55</td>
<td>10.31</td>
<td>8.48</td>
<td>X</td>
</tr>
</tbody>
</table>

Figure 4.1 LPDA with labeled dimensions
This structure was created in HFSS and simulated over the frequency range 0.7 – 6.0 GHz. The model is seen in Figure 4.2. The feed line enters the structure from the left and is delivered to the dipole array through a 50 Ω coaxial line. The center conductor of the coaxial cable is connected to the pipe of similar radius at the bottom of the structure. This allows for the interweaving arm connections required for the LPDA. The spacing between the coaxial cable and pipe also satisfy the dimensions for a 50 Ω balanced line. The dimensions are listed in Table 4.3.

Figure 4.2 HFSS model of LPDA
The VSWR of the antenna and the normalized realized gain patterns of selected frequencies are shown in Figure 4.3 and Figure 4.4 (a)-(l). The VSWR is not typically less than 2:1, but rather 3.5:1 for most of the design band. The average realized gain of Figure 4.4 (a)-(l) was 5.78 dBi with an average front-to-back ratio of 12.02 dB. The realized gain is 0.78 dB higher than the typical gain given on the HyperLOG® 7060 datasheet and the front-to-back ratio is estimated to be 7 dB less than the datasheet’s figures.

![Figure 4.3 Simulated VSWR of LPDA](image-url)
Figure 4.4 Normalized realized gain of LPDA, $\theta = 90^\circ$ plane at (a) 700 MHz (b) 1.0 GHz (c) 1.5 GHz (d) 2.0 GHz (e) 2.5 GHz (f) 3.0 GHz (g) 3.5 GHz (h) 4.0 GHz (i) 4.5 GHz (j) 5.0 GHz (k) 5.5 GHz and (l) 6.0 GHz
4.2. SIMULATED LPDA AGAINST MATERIAL SLAB

A material slab was placed in the direction of the main beam of the LPDA, centered on the axis of EM radiation. This slab was 500 mm x 500 mm x 0.1 mm and assigned the conductivity and relative permeability determined earlier for the alloy material. The variant without surface roughness was used to increase the simulation speed. Figure 4.5 shows how the material was placed in front of the array. Four distances were examined to determine an optimal separation between the tip of the array and the material slab: 100 mm, 75 mm, 50 mm, and 25 mm. The input power was maintained at 1 W for all four trials.

At the completion of each simulated distance, the current density on the material can be visualized, as shown in Figure 4.6. The current density does vary with the phase of the input signal and the magnitude is dependent on the separation. Figure 4.6 was animated with the input signal phase and then stopped when the current density contained a maximum at the center of the slab. The separation was 100 mm for these images.

![Figure 4.5 Material slab separated 100 mm from the LPDA](image-url)
4.3. Determining Rise in Temperature with Simulated LPDA Exposure

In order to calculate the rise in temperature of the alloy material when exposed to the EM radiation, the current density on the material slab was exported. This is done by placing 42 lines on the slab, starting at the left edge. The field calculator in HFSS can create a table of current densities along the lines. 42 points are created on each line, resulting in a square matrix of current densities on the slab. Because the values are from a line, the table is in units of A/m. These values are brought into MATLAB and the perimeter of the matrix is ignored to avoid any anomalies at the edges, leaving a 40 x 40 matrix. This pruning technique is described in Figure 4.7. This figure shows the top left corner of the material slab during exposure. The solid lines are the non-model objects that HFSS uses.
Figure 4.7 Top left corner of material slab, showing where current density is measured to extract the current densities. The solid line intersections convey where this measurement is extracted along the line. By removing the outer edge of the matrix, the dashed boxes are created, with the measurements taken in the center of each box. The measurement is assumed to represent the average current located inside of each dashed box.

For each of the four distances, the current density matrix was brought into MATLAB and the edge pruning technique was performed during the import. Each element is multiplied by the distance between elements, creating a matrix of currents. From here, each element is squared and multiplied by the surface resistivity $R_s$. The resulting matrix is the power absorbed by the material. Each element can then be multiplied by an exposure time to yield the energy absorbed over the timeframe. Lastly, the specific heat and mass of the material is utilized to determine the change in temperature in each element. At each
import J_matrix;
unitDistance = materialSlabLength / (numLines-1);
I_matrix = J_matrix * unitDistance;
P_matrix = I_matrix.^2 * Rs;
E_matrix = P_matrix * exposureTime;
unitMass = density * slabVolume / numElements;
dT_matrix = E_matrix / specificHeat / unitMass;

Figure 4.8 MATLAB script pseudocode for determining rise in temperature

stage, the matrix is synonymous to the average value inside of each dashed box in Figure 4.7.

The unit distance used to create the current matrix is 12.195 mm, the side length of each element. The surface resistivity is 1.021 Ω and is determined from the conductivity and relative permeability defined in Chapter 3. The density of the alloy material was also determined during the works of Chapter 3 by measuring the mass of material before it was compressed into the known groove volume of the FR4 boards. The alloy’s density is 5444 kg/m³. The unit mass is 0.085 g and can be calculated with the known density, slab volume, and number of elements.

The specific heat capacity of the alloy material is unknown. This term is a critical component in determining the rise in temperature of the material and can only be estimated at this point. Some chemical element composition of the alloy is known and the specific heat was estimated from this data. The estimation was based on a weighted combination of the two primary elements in the alloy, yielding 0.257 J/kg/K. It is important to note that a weighted combination of elemental composition is not an appropriate method for determining or estimating the specific heat capacity of a material. This value is extraneous and may not lead to confident results, which will be explored and discussed later. An accepted method for determining this value experimentally is through the use of a bomb
Table 4.4 Calculated Values for Alloy Slab near LPDA after 5 Minute Exposure

<table>
<thead>
<tr>
<th>Separation</th>
<th>Slab Total Power</th>
<th>Slab Avg. ΔT**</th>
<th>Hot Spot* Power</th>
<th>Hot Spot* ΔT**</th>
</tr>
</thead>
<tbody>
<tr>
<td>100 mm</td>
<td>72.87 mW</td>
<td>0.62 °C</td>
<td>0.45 mW</td>
<td>6.20 °C</td>
</tr>
<tr>
<td>75 mm</td>
<td>65.17 mW</td>
<td>0.56 °C</td>
<td>0.47 mW</td>
<td>6.38 °C</td>
</tr>
<tr>
<td>50 mm</td>
<td>54.24 mW</td>
<td>0.47 °C</td>
<td>0.60 mW</td>
<td>8.22 °C</td>
</tr>
<tr>
<td>25 mm</td>
<td>114.19 mW</td>
<td>0.99 °C</td>
<td>2.05 mW</td>
<td>28.04 °C</td>
</tr>
</tbody>
</table>

* Hot spot is the average value of a 0.5 inch diameter were most power is absorbed

** Rise in temperature value is based on extraneous value for specific heat capacity calorimeter which is unavailable. An attempt at refining the specific heat capacity for the alloy is presented in Chapter 5.

Table 4.4 shows the results of calculating the rise in temperature when using this current density method. The separation of the material from the tip of the array is given along with the quantities for the total slab and a defined hot spot. The hot spot on the material is a circle with diameter of 0.5 inch where the highest absorbed power occurs. The total slab power slightly decreases as the separation decreases until the closest distance. This trend was not expected, however is attributed to the radiation pattern deformation as the material is brought closer. At the closest distance, the power should increase solely because of the proximity. The hot spot does follow an increasing power trend as the distance is reduced, suggesting that less power is absorbed near the edges as the antenna moves closer, which also explains the decreasing power trend with the total slab. It is important to note that the temperature calculation is derived from the estimated specific heat capacity, which could not accurately represent the material.
4.4. ISSUES WITH LPDA

The LPDA in free space exhibits excellent qualities of an antenna that would be used in a traditional communications aspect. The desired high directivity does not have room to materialize when the alloy material is close to the array. The most active element on the array when operating at 2.4 GHz is the 7th set of arms, which is 65.17 mm from the tip. When the slab is 25 mm from the tip, this active element is 90.17 mm from the slab. This distance is considered to be too great for significant energy transfer, suggesting that the LPDA may not be the best means of an applicator. Furthermore, Figure 4.9 demonstrates the slab has nearly reflected all the EM radiation. The material is in the $\phi = +90^\circ$ direction where very little power is available.

![Graph of normalized realized gain of LPDA with 25 mm separation, slab in $\phi = +90^\circ$ direction](image)

Figure 4.9 Normalized realized gain of LPDA with 25 mm separation, slab in $\phi = +90^\circ$ direction

In conclusion, the LPDA is not useful as an applicator in this situation. The directivity of the array is destroyed in the close presence of the material and the active element is too far from the material. The antenna could be more effective if the higher
frequency element were used, as it is closer to the tip of the device and the material. A dipole antenna will be analyzed next because it can be constructed to operate within the allocated frequency band and move much closer to the material in order to increase energy transfer.
CHAPTER 5. SIMULATED ENERGY TRANSFER WITH HALF-WAVELENGTH DIPOLE AND EXPERIMENTAL RESULTS

If bringing an antenna close to the material continues to increase the amount of power absorbed, an antenna must be selected where the radiating element is allowed to move as close as possible to the material. To simplify the experimental design and overall heating process, a half wavelength dipole is considered to transfer energy to the alloy material at 2.4 GHz. The log-periodic dipole array’s active element near this frequency range is too far from the material to give considerable heating, whereas the dipole can be brought much closer to increase the absorbed power. This configuration also allows for a smaller application device and overall lower profile. A dipole was constructed with a bazooka-style balun in the Microwave Engineering Laboratory and sent to the Savannah River National Laboratory for experimental measurements along with a piece of thin graphite. The same antenna was then modeled in HFSS with a piece of graphite in order to understand and calibrate the heating calculations between the simulated results and experimental measurements. With all of the characteristics of graphite known, such as the specific heat capacity, this calibration process enables more confidence when the fabricated dipole is used in the same configuration with the alloy material. The heating results for both the graphite and alloy sheets are presented here.
5.1. DIPOLE ANTENNA DESIGN

The dipole was designed to operate at 2.4 GHz with a quarter wavelength bazooka balun to transition from the unbalanced coaxial feed. The HFSS model of this structure is shown in Figure 5.1 followed by the fabricated device in Figure 5.2.

One half a wavelength and one quarter a wavelength at 2.4 GHz is 62.5 mm and 31.25 mm respectively, however HFSS was allowed to optimize the dipole and balun lengths to give the best response at the design frequency. This yielded lengths of 59.162 mm and 29.162 mm. The fabricated dipole was made to these dimensions and trimmed while connected to a vector network analyzer. The dipole arms were shortened until an appropriate $S_{11}$ was achieved on the VNA at 2.4 GHz. The final dimensions for the dipole are illustrated in Figure 5.2. The balun was constructed from a piece of 0.5 inch diameter copper pipe. The VNA measurement is shown below in Figure 5.3. There exists a good match at 1.547 GHz and 2.403 GHz with $S_{11}$ at -17.22 dB and -18.81 dB respectively.

![Figure 5.1 HFSS model of dipole antenna with balun](image-url)
Figure 5.2 Fabricated dipole antenna with balun

Figure 5.3 $S_{11}$ of fabricated dipole measured by VNA
5.2. SIMULATED DIPOLE ANTENNA AGAINST GRAPHITE SLAB

The dipole was then modeled above the surface of a graphite sheet as seen in Figure 5.4. The graphite is PGS (Pyrolytic Graphite Sheet) from Panasonic with part number EYGS121810. The material parameters are listed in Table 3.1.

![Figure 5.4 HFSS model of dipole above material slab showing non-model lines for current density extraction](image)

<table>
<thead>
<tr>
<th>Table 5.1 Panasonic PGS EYGS121810 Specifications</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Width</strong></td>
</tr>
<tr>
<td><strong>Length</strong></td>
</tr>
<tr>
<td><strong>Thickness</strong></td>
</tr>
<tr>
<td><strong>Conductivity</strong></td>
</tr>
<tr>
<td><strong>Relative Permeability</strong></td>
</tr>
<tr>
<td><strong>Specific Heat Capacity</strong></td>
</tr>
<tr>
<td><strong>Density</strong></td>
</tr>
</tbody>
</table>
The graphite sheet is used as an experimental benchmark because it is a well-documented material. The electrical properties of the material are known as well as the specific heat, which is required to determine the rise in temperature when energy is transferred.

42 non-model lines are placed on the surface of the material in a similar fashion and for the same purpose as described in Section 4.3. The outer parameter is ignored to focus on the center 40 x 40 matrix of current density. The same MATLAB post-processing technique is performed as described in Figure 4.8, however the values of some of the variables have changed with the new material and physical dimensions. Unlike the LPDA simulations, the distance from the graphite is not varied but held constant at 5/16 inch between the surface of the graphite and dipole arms. The input power was selected to change for each trial and varied from 5 W, 10 W, and 20 W.

<table>
<thead>
<tr>
<th>Input Power</th>
<th>Slab Total Power</th>
<th>Slab Avg. ΔT</th>
<th>Hot Spot* Power</th>
<th>Hot Spot* ΔT</th>
</tr>
</thead>
<tbody>
<tr>
<td>5 W</td>
<td>29.21 mW</td>
<td>9.2 °C</td>
<td>6.69 mW</td>
<td>224.2 °C</td>
</tr>
<tr>
<td>10 W</td>
<td>58.41 mW</td>
<td>18.3 °C</td>
<td>13.39 mW</td>
<td>448.3 °C</td>
</tr>
<tr>
<td>20 W</td>
<td>116.82 mW</td>
<td>36.7 °C</td>
<td>26.77 mW</td>
<td>896.6 °C</td>
</tr>
</tbody>
</table>

* Hot spot is the average value of a 0.5 inch diameter were most power is absorbed

The above table is calculated using the method described in Section 4.3. This method is flawed in that it does not consider how the heat is conducted in the material, meaning that energy does not flow out of each matrix element, which is incorrect. There is a considerable amount of surface area that would radiate the heat further into the environment. Because of the inaccuracy of this model, the calculated values here will be extraneous, but a scaling technique can be applied as a solution.
5.3. EXPERIMENT WITH FABRICATED DIPOLE ANTENNA AGAINST GRAPHITE SLAB

The values in Table 5.2 assume that the transferred energy does not leave the matrix element into other elements or into the air. To compensate and calibrate this methodology, a scaling factor can be applied to the temperature calculation if experimental data is analyzed. The fabricated dipole and the graphite sheet were used to perform these measurements. The dipole was connected to a VNA through a power amplifier with a three point sweep range starting at 2.4 GHz and a long enough sweep step time to record the temperature. This type of setup on the VNA enables it to imitate a signal generator. The VNA was able to provide the signal with output power of -9 dBm, -6 dBm, and -3 dBm for three different trials. The power amplifier has a measured gain of 49.04 dB at 2.4 GHz. The cable loss in the experiment was measured to be 2.74 dB at 2.4 GHz. This setup results in 5.05 W, 10.72 W, and 21.38 W available to the dipole.

Figure 5.5 Schematic of setup for measuring experimental rise in temperature using a VNA, power amplifier, and dipole antenna
Figure 5.6 Experimental setup with dipole exposing PGS in presence of an IR camera. Courtesy of Michael Brown, Savannah River National Laboratory

The experiment is shown Figure 5.6. The dipole was suspended 5/16 inch above the surface of the graphite. A foam table was constructed to hold the antenna and graphite because the foam will not introduce interference with the measurements. The process was conducted inside of a closed metallic chamber with the VNA and power amplifier outside. The material was exposed to each of the three input power levels and allowed ample time to cool between each trial. The infrared camera captured the temperature contours on the material during each trial and was used to record the temperature every 60 seconds during the exposure periods. Figure 5.7 is an image taken by the infrared camera at the end of the 20 W input trial. This image shows that the hot spot on the material is approximately the diameter of the dipole’s balun: 0.5 inch.
Figure 5.7 Image obtained from infrared camera showing the 0.5 inch diameter hot spot. Courtesy of Michael Brown, Savannah River National Laboratory

Figure 5.8 Plot of PGS temperature in the hot spot for all trials. Courtesy of Michael Brown, Savannah River National Laboratory
Table 5.3 Measured Hot Spot Temperature Values for PGS Slab near Dipole (°C)

<table>
<thead>
<tr>
<th>Input Power</th>
<th>Start</th>
<th>1 min</th>
<th>2 min</th>
<th>3 min</th>
<th>4 min</th>
<th>5 min</th>
</tr>
</thead>
<tbody>
<tr>
<td>5 W</td>
<td>20.3</td>
<td>22.0</td>
<td>22.7</td>
<td>23.3</td>
<td>23.7</td>
<td>23.9</td>
</tr>
<tr>
<td>10 W</td>
<td>20.3</td>
<td>23.4</td>
<td>25.3</td>
<td>26.5</td>
<td>27.4</td>
<td>28.2</td>
</tr>
<tr>
<td>20 W</td>
<td>20.3</td>
<td>28.8</td>
<td>31.3</td>
<td>33.5</td>
<td>36.8</td>
<td>38.0</td>
</tr>
</tbody>
</table>

The above table contains the data used to generate plot in Figure 5.8. These data points were measured using the infrared camera. Each of the three trials begin to approach a temperature asymptote, suggesting an equilibrium temperature during each trial. This type of equilibrium will occur as the material begins to sink heat away from the hotspot as the dipole continues to provide constant power. When the energy being absorbed into the material is equal to the energy dissipated into the environment, this moment is the temperature equilibrium.

5.4. COMPARING SIMULATED AND EXPERIMENTAL GRAPHITE TEMPERATURE

The simulated and experimental results can be compared to determine an appropriate scaling factor to compensate for the heat energy that is conducted away from the hot spot. This efficiency factor will be used to consume the undetermined physics behind the thermal conductivity of the material, although the thermal conductivity is known. This term is useful because it allows us to place all dependence of heat transfer in the material and the thermal dissipation into the air in to one approximated term. The experimental heating measurements demonstrate what will actually happen at the hotspot when heat is allowed to flow away from the center and dissipate into the environment, whereas the analytical approach by means of the specific heat capacity alone does not allow this.
During the experiment, all initial temperatures were 20.3°C and increased to 23.9°C, 28.2°C, and 38.0°C for each of the three trials, respectively. The experimental ΔT is shown in Table 5.4. An Efficiency Factor (EF) is introduced into the calculation to reduce the previous ΔT values calculated in Table 5.2. Because the actual heating is nonlinear with an asymptote and the MATLAB calculation is strictly linear, this value only satisfies the change in temperature for this material at the 5 minute exposure, shown in Figure 5.9.

Table 5.4 Adjusted ΔT for PGS Slab near Dipole after 5 Minute Exposure

<table>
<thead>
<tr>
<th>Input Power</th>
<th>Hot Spot* ΔT unaltered**</th>
<th>Hot Spot* ΔT with E.F.</th>
<th>Measured ΔT</th>
</tr>
</thead>
<tbody>
<tr>
<td>5 W</td>
<td>224.2 °C</td>
<td>4.0 °C</td>
<td>3.6 °C</td>
</tr>
<tr>
<td>10 W</td>
<td>448.3 °C</td>
<td>7.9 °C</td>
<td>7.9 °C</td>
</tr>
<tr>
<td>20 W</td>
<td>896.6 °C</td>
<td>15.9 °C</td>
<td>17.3 °C</td>
</tr>
</tbody>
</table>

* Hot spot is the average value of a 0.5 inch diameter were most power is absorbed
** Calculation does not consider thermal conductivity
E.F. is efficiency factor of 0.0177

Figure 5.9 EF fitting at 10 W trial between experiment and linear calculation
5.5. **Comparing Simulated and Experimental Alloy Temperature**

The model displayed in Figure 5.4 was used again with the alloy material substituted in as the material slab. This only requires changing the conductivity and relative permeability of the model and boundary to the parameters defined in Chapter 3. The values used are those without the surface roughness to increase the simulation speed. The non-model lines were used to extract the current density and perform the same analytical calculations as before to determine the rise in temperature. The calculations are shown in Table 5.5. Again, these values do not consider how the heat is carried away from the hot spot and will appear to be large.

<table>
<thead>
<tr>
<th>Input Power</th>
<th>Slab Total Power</th>
<th>Slab Avg. ΔT**</th>
<th>Hot Spot* Power</th>
<th>Hot Spot* ΔT**</th>
</tr>
</thead>
<tbody>
<tr>
<td>5 W</td>
<td>226.11 mW</td>
<td>36.6 °C</td>
<td>51.67 mW</td>
<td>892.6 °C</td>
</tr>
<tr>
<td>10 W</td>
<td>452.23 mW</td>
<td>73.2 °C</td>
<td>103.34 mW</td>
<td>1785.2 °C</td>
</tr>
<tr>
<td>20 W</td>
<td>904.45 mW</td>
<td>146.5 °C</td>
<td>206.67 mW</td>
<td>3570.5 °C</td>
</tr>
</tbody>
</table>

*Hot spot is the average value of a 0.5 inch diameter were most power is absorbed
**Rise in temperature value is based on extraneous value for specific heat capacity and does not consider thermal conductivity

The same experimental setup was performed with the alloy material inside of the metallic chamber to collect its hot spot temperature after being exposed to the dipole for 5 minutes. This is to provide the same calibration method to correct for the unrealistic temperature increases in Table 5.5. The alloy material also has unknown properties with how the heat is dissipated in the material. This material also is granular and will not conduct heat away as well as the PGS material. The grains can be seen in Figure 5.10. These experimental results are shown in Figure 5.11.
Figure 5.10 Experimental setup with dipole exposing alloy. Courtesy of Michael Brown, Savanah River National Laboratory

Figure 5.11 Plot of alloy temperature in the hot spot for all trials. Courtesy of Michael Brown, Savanah River National Laboratory
There is still a discrepancy between the experimental measurements and the calculated values. The values can be forced to converge in a similar manner as before by applying an efficiency factor to the calculated change in temperature until it matches the measured change in temperature. Physically, this includes the thermal conductivity of the material, how the heat is transferred in the material, and how effectively the heat is then radiated into the environment. A spacing factor term is also included for this granular material, which the PGS slab does not have. This process is shown in Table 5.7.

### Table 5.6 Measured Hot Spot Temperature Values for Alloy Slab near Dipole (°C)

<table>
<thead>
<tr>
<th>Input Power</th>
<th>Start</th>
<th>1 min</th>
<th>2 min</th>
<th>3 min</th>
<th>4 min</th>
<th>5 min</th>
</tr>
</thead>
<tbody>
<tr>
<td>5 W</td>
<td>21.0</td>
<td>22.7</td>
<td>23.4</td>
<td>23.6</td>
<td>23.7</td>
<td>23.9</td>
</tr>
<tr>
<td>10 W</td>
<td>20.6</td>
<td>23.0</td>
<td>24.2</td>
<td>24.6</td>
<td>25.0</td>
<td>25.1</td>
</tr>
<tr>
<td>20 W</td>
<td>20.6</td>
<td>26.1</td>
<td>28.1</td>
<td>29.3</td>
<td>29.8</td>
<td>30</td>
</tr>
</tbody>
</table>

### Table 5.7 Adjusted ΔT for Alloy Slab near Dipole after 5 Minute Exposure

<table>
<thead>
<tr>
<th>Input Power</th>
<th>Hot Spot* ΔT unaltered**</th>
<th>Hot Spot* ΔT with E.F. + S.F.</th>
<th>Measured ΔT</th>
</tr>
</thead>
<tbody>
<tr>
<td>5 W</td>
<td>892.6 °C</td>
<td>2.3 °C</td>
<td>2.9 °C</td>
</tr>
<tr>
<td>10 W</td>
<td>1785.2 °C</td>
<td>4.6 °C</td>
<td>4.5 °C</td>
</tr>
<tr>
<td>20 W</td>
<td>3570.5 °C</td>
<td>9.3 °C</td>
<td>9.4 °C</td>
</tr>
</tbody>
</table>

* Hot spot is the average value of a 0.5 inch diameter were most power is absorbed
** Rise in temperature value is based on extraneous value for specific heat capacity and does not consider thermal conductivity
E.F. is efficiency factor of 0.0177
S.F. is spacing factor of 0.1470

In conclusion, the heating calculations for both materials are adjusted by an efficiency factor to account for the unknown thermal properties of the materials. This must be done because the calculated values are much larger than the experimental results. It was
determined by the experiment that this occurs because the calculation method does not account for the thermal conductivity of either material, thus assuming that all energy stays where it is absorbed and does not radiate into the environment or across the materials. These scaling values are relatively small and suggest that the transfer of heat across the material is an essential property in the calculations. The PGS material will have greater heat dissipation because it is a solid material, unlike the alloy which is granular and unconnected. Now that a method has been found to approximate the rise in temperature in a slab of the material, the alloy material will be arranged into a structure as to increase the energy transfer at a selected frequency.
CHAPTER 6. INCREASING ENERGY TRANSFER WITH A CORRUGATED SPLIT RING RESONATOR

The alloy material will absorb some of the RF energy when it is placed close to the dipole antenna. This amount is more significant than the exposure near the LPDA because of the distance from the active element, however the material could be fashioned into a split ring resonator to further increase the amount of energy it absorbs. Traditionally, split ring resonators (SRR) are placed together to form a frequency selective surface (FSS) that utilize a physical shape to modify an antenna’s radiation pattern. This chapter will discuss how a custom SRR can be created with the material to increase the material heating.

6.1. CORRUGATED SPLIT RING RESONATOR SELECTION

The selected SRR was designed for use around 2.4 GHz [33]. This corrugated design is a modified ring resonator in that the layout is a square with multiple teeth. One traditional design is shown in Figure 6.1. This figure has dimensions labeled to illustrate the variables that could change the resonator. The split ring acts as an LC pair, the ring structure creating an inductance and the gap in the ring creating a capacitance. Figure 6.2 contains the corrugated design. Inductance is from the perimeter path of the material whereas the capacitance comes from the gap created by the teeth. The corrugated design shall be utilized for increasing the energy transfer. The dimensions shown in Figure 6.2 are extracted from the reference to operate at 2.4 GHz. The SRR is placed on top of a 1.6 mm thick FR4 substrate with a relative permittivity of 4.4 and side lengths of 18 mm [33].
Figure 6.1 Traditional double split ring resonator with labeled variables

Figure 6.2 Corrugated split ring resonator for 2.4 GHz with labeled dimensions. Extracted from [33]
6.2. Split Ring Resonator Validation

The SRR model was created in HFSS, as shown in Figure 6.2. The material was assigned the graphite PGS definition. Six cases were tried to see if the SRR had higher induced currents depending on where an antenna was positioned around the structure. A single 2.4 GHz half wavelength dipole was positioned on all six sides for the six different trials. These locations were setup such that the antenna was 5/16 inch from the closest point of the SRR. Figure 6.3 explains these orientations with the assistance from a cube model. If the SRR is thought of as a cube, the antenna was placed on all six sides of the cube for the six different trials. For each trial, the broad side of the dipole was parallel to the SRR so that the antenna’s null was not fixated toward the resonator.

Because of symmetry, the six cases can be simplified into five trials: Left, Right, Top, Bottom, and Front. The Front and Back sides shall have the same magnitude of induced current because of a mirrored axis.

The current density contours in the flowing figures show the highest magnitude when the antenna is positioned on the Left. This implies that the most optimal side for the dipole antenna is on the Left side, with the capacitive teeth.

Figure 6.3 3D model of SRR and reference cube with labeled directional sides
Figure 6.4 Surface current density contours when dipole is on left of SRR

Figure 6.5 Surface current density contours when dipole is on right of SRR
Figure 6.6 Surface current density contours when dipole is at front or back of SRR

Figure 6.7 Surface current density contours when dipole is on top of SRR
Figure 6.8 Surface current density contours when dipole is on bottom of SRR

Figure 6.9 Full view of corrugated SRR HFSS model with 2.4 GHz dipole positioned 5/16 inch on Left
A frequency sweep was also simulated near the resonator to determine how responsive the structure is to the design frequency. The sweep range was 1.88-3.78 GHz with 20 points. The dipole length changed at each data point so that the antenna design was optimal for the recording. Each S$_{11}$ value was at most -15 dB at the recorded frequency.

| Table 6.1 Recorded Maximum J$_{surf}$ During Frequency Sweep on PGS |
|-------------------------|---------|---------|---------|---------|---------|---------|---------|---------|---------|
| Freq (GHz)              | 1.88    | 1.93    | 1.98    | 2.05    | 2.09    | 2.16    | 2.23    | 2.29    | 2.37    | 2.45    |
| Dipole (mm)             | 37      | 36      | 35      | 34      | 33      | 32      | 31      | 30      | 29      | 28      |
| J$_{surf}$ (A/m)        | 15.3    | 11.6    | 11.0    | 15.1    | 15.0    | 17.0    | 24.0    | 25.3    | 19.8    | 18.6    |
| Freq (GHz)              | 2.56    | 2.62    | 2.70    | 2.78    | 3.01    | 3.14    | 3.27    | 3.44    | 3.59    | 3.79    |
| Dipole (mm)             | 27      | 26      | 25      | 24      | 23      | 22      | 21      | 20      | 19      | 18      |
| J$_{surf}$ (A/m)        | 15.9    | 14.6    | 16.2    | 16.6    | 13.1    | 13.2    | 12.9    | 11.3    | 12.5    | 13.3    |

Figure 6.10 Plot of magnitude of maximum surface current density on PGS corrugated SRR during frequency sweep
The frequency sweep revealed that the corrugated SRR has a current density maximum slightly below the design frequency at 2.29 GHz. This maximum is approximately 1.39 times greater than the maximum that occurs at the design frequency.

These results suggest that the corrugated split ring resonator will have the most induced current on its surface when the dipole antenna is tuned and operating at 2.29 GHz on the side of the structure with the capacitive teeth.

6.3. Comparing PGS Corrugated SRR with Equivalent Slab

In order to verify that the corrugated SRR does perform better at heating than a slab of the material, a HFSS model of a slab on the same FR4 substrate was created. This square slab shared a similar side length of 10.125 mm and was held at a distance of 5/16 inch between the edge of the slab and the dipole antenna. The dipole was operated at 2.29 GHz in order to generate the current density contour in Figure 6.11.

Figure 6.11 Surface current density contour on PGS slab at 2.29 GHz
The PGS slab of equal size contained current densities on its surface that peaked at 8.0732 A/m with a significant portion of the area being less than 5.0 A/m. The corrugated split ring resonator was able to induce a maximum current density of 25.3007 A/m, 3.12 times greater than the slab.

6.4. Comparing Experimental Alloy Corrugated SRR with Equivalent Slab

Finally, the two HFSS models for the corrugated split ring resonator and slab were used again with the boundary definition of the alloy metal parameters without the surface roughness, defined in Chapter 3. Figure 6.12 demonstrates that the alloy material, when fabricated as the resonator, can achieve a maximum surface current density of 11.7228 A/m. With the alloy slab alone, as in Figure 6.13, the maximum surface current density is 5.6031 A/m. With the use of the resonator, the maximum surface current density is 2.09 times greater than the slab for the alloy material.

Figure 6.12 Surface current density contour on alloy corrugated SRR at 2.29 GHz
6.5. DETERMINING TEMPERATURE INCREASE ON CORRUGATED SPLIT RING RESONATOR

Regardless of the material, the corrugated split ring resonator increases the maximum induced surface current density in both materials. These increases are enough reason to further analyze how much power is absorbed by the resonator to estimate the temperature increase. The same current density method summarized in Figure 4.8 is used to estimate the temperature increase.

First, the current densities along the PGS slab and corrugated split ring resonator were exported to MATLAB for the post-processing technique. The current density values were obtained by placing non-model lines in the center of each of the capacitive teeth and around the square ring for the resonator. The data on each line was exported every 0.5 mm
Figure 6.14 Corrugated split ring resonator divided into 0.5 mm square units

Figure 6.15 Slab divided into 0.5 mm square units
so that each point represented a square unit on the ring. This value was chosen because the narrowest part of the SRR is 0.5 mm, thus creating the unit squares. This idea is conveyed graphically in Figure 6.14. The current density data for the PGS slab was also imported for every 0.5 mm square, shown in Figure 6.15. The results are shown in Table 6.2. These results are the total increase in temperature after a 5 minute exposure with three different input power values to the dipole. These results were obtained by the extracted surface current densities but were scaled to the previous results from Chapter 5 because of the similar areas between the control slab and the hot spot. In the previous case, the experimental validation was able to provide an efficiency factor to scale the results. The area of the hot spot was 126.67 mm² whereas the area of this slab is 102.52 mm². The calculations for the slab here are normalized to fit these previous values by a factor of 3,509. This forces a fit between this slab and the previously determined rise in temperature. This same calculation is performed for the SRR and demonstrates how effective the SRR is with the rise in temperature. The ratio of temperature increase between each input power trial for the PGS SRR and slab is a factor of 5.6354 for the calculations.

<table>
<thead>
<tr>
<th>Structure</th>
<th>Input Power</th>
<th>ΔT unaltered*</th>
<th>ΔT adjusted**</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slab</td>
<td>5 W</td>
<td>13,947.4 °C</td>
<td>4.0 °C</td>
</tr>
<tr>
<td>SRR</td>
<td>5 W</td>
<td>78,627.1 °C</td>
<td>22.4 °C</td>
</tr>
<tr>
<td>Slab</td>
<td>10 W</td>
<td>27,894.8 °C</td>
<td>7.9 °C</td>
</tr>
<tr>
<td>SRR</td>
<td>10 W</td>
<td>157,254.2 °C</td>
<td>44.8 °C</td>
</tr>
<tr>
<td>Slab</td>
<td>20 W</td>
<td>55,789.7 °C</td>
<td>15.9 °C</td>
</tr>
<tr>
<td>SRR</td>
<td>20 W</td>
<td>314,508.4 °C</td>
<td>89.6 °C</td>
</tr>
</tbody>
</table>

*Calculation does not consider thermal conductivity
** Adjustment occurs by normalizing the slab results to the previous hotspot results and is accomplished by removing a factor of 3,509
Table 6.3 contains the temperature increase values for a corrugated split ring resonator and slab when fabricated from the alloy material. As with the PGS material, the resulting temperature calculation is scaled by a factor of 26,262 to force a normalization between the current slab structure and the previous hot spot from Chapter 5.

The resonator can induce more current in the alloy corrugated split ring resonator than an equivalent sized slab leading to a 5.5019 ratio of temperature increase at each input power trial.

<table>
<thead>
<tr>
<th>Structure</th>
<th>Input Power</th>
<th>(\Delta T) unaltered*</th>
<th>(\Delta T) with E.F. + S.F.</th>
</tr>
</thead>
<tbody>
<tr>
<td>Slab</td>
<td>5 W</td>
<td>60,882.5 °C</td>
<td>2.3 °C</td>
</tr>
<tr>
<td>SRR</td>
<td>5 W</td>
<td>333,339.9 °C</td>
<td>12.7 °C</td>
</tr>
<tr>
<td>Slab</td>
<td>10 W</td>
<td>121,764.9 °C</td>
<td>4.6 °C</td>
</tr>
<tr>
<td>SRR</td>
<td>10 W</td>
<td>666,679.8 °C</td>
<td>25.4 °C</td>
</tr>
<tr>
<td>Slab</td>
<td>20 W</td>
<td>243,529.9 °C</td>
<td>9.3 °C</td>
</tr>
<tr>
<td>SRR</td>
<td>20 W</td>
<td>1,333,359.7 °C</td>
<td>50.8 °C</td>
</tr>
</tbody>
</table>

* Rise in temperature value is based on extraneous value for specific heat capacity and does not consider thermal conductivity

** Adjustment occurs by normalizing the slab results to the previous hotspot results and is accomplished by removing a factor of 26,262
CHAPTER 7. CONCLUSION AND FUTURE WORK

7.1. CONCLUSION

In this thesis a methodology was developed to estimate the conductivity and relative permeability of a granular metal-magnetic alloy material. The alloy was used to develop a microstrip transmission line by placing and compacting the granular material into a narrow groove that was milled into an FR4 Printed Circuit Board. Simulations were performed to de-embed the losses due to the FR4 dielectric material. By comparing the simulated and measured S-parameter results, especially once the mismatch and dielectric losses were removed from the calculation, the alloy’s conductivity and relative permeability could be estimated. It was found that the ratio between the two is the governing parameter. This ratio was $1 \times 10^6$ by taking into account for the surface roughness of the material while it was $2.5 \times 10^5$ without any surface roughness. The latter was considered to significantly reduce the simulation times with specimens exposed to the near fields of antennas. The above estimated conductivity/relative permeability ratio was then used in all forthcoming simulation models to investigate and understand the RF energy absorbed by the alloy material in the presence of antennas.

To fully understand and calibrate the conversion of absorbed RF energy into heat, given that not 100% of the RF energy absorbed results in temperature increase in the material, a second methodology was developed. This methodology is used to estimate the temperature increase in a known material, graphite, when it was exposed to the near field
EM radiation from a resonant dipole antenna operating at 2.4 GHz. Comparing the simulated temperature rise with experimentally measured temperature increase in graphite for various input powers to the antenna an Efficiency Factor (EF) of 1.77% was determined. For example, it was found that for 10 W of input power to the antenna with a graphite separation distance of 5/8 inch, the temperature increase for 5 minutes of exposure was 7.9 °C. The implication of the EF is that it is quite challenging to determine the heat loss due to convection or other mechanisms and that the measured temperature increase is significantly lower compared to the simulated temperature rise. These observations warrant an EF to fully calibrate the simulated results. Similarly, for the granular alloy material the experimental temperature rise for 5 minutes of RF exposure from 10 W input power fed to a resonant dipole at 2.4 GHz resulted in 4.5 °C temperature rise. To account for the granular nature and the material grains being quite dispersed from each other in many areas a Spacing Factor (SF) was introduced into estimating the simulated temperature rise. The SF considered was 14.7%.

Finally, a new method of further enhancing the EM energy absorption in material was developed by designing and investigating a Split Ring Resonator (SRR) in simulations which showed the potential to increase the temperature rise in the alloy by a factor of 5.5 due to its inherently resonant nature.

7.2. FUTURE WORKS

Future work should include an experimental measurement of the resistivity of the alloy material to deduce the value of the relative permeability. If the resistivity is measured on a fabricated piece of the alloy as a control sample, the ratio of relative permeability and conductivity determined for the estimated surface roughness can be used to calculate this
value. If possible, the calculated relative permeability should be cross referenced with another instrument that can measure this quantity to determine the reliability of this process.

Furthermore, if the input power of the LPDA were instead applied at 6.0 GHz, the active element would be the closest to the slab, increasing the induced current on the material. Based on the surface resistance calculation, there should be a higher surface resistance at this frequency, which could lead to high absorbed power if the current density does not change. This idea should be simulated and verified experimentally though. This thought should reveal more about how frequency plays a role in the heating of the material and the temperature contours. It has been noted with antennas in tissue that an increasing frequency will cause a smaller area to heat, but the power increase from the high frequency results in more losses with thermal conductivity aiding in the increase of the hot spot.

Finally, there should be continued experimental validation of the temperature calculations with the corrugated split ring resonator. It is reasonable to scale the results with the 102.52 mm² slab by the 126.67 mm² spot to adjust the heating calculation for the resonator, but it should not be taken for granted. This calculation is enough ground to demonstrate that the resonator will increase the heating significantly, but the SRR should be fabricated in order to measure the actual increase in temperature.
REFERENCE


