Metamaterial Enhanced Wireless Power Transmission System

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Master of Science in Electrical Engineering

by
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ABSTRACT

Metamaterial Enhanced Wireless Power Transmission System

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Nikolai Tesla's revolutionary experiments demonstrated the possible benefits of transmitting power wirelessly as early as 1891. Applications for the military, consumers, emergency personnel, remote sensors, and others use Tesla's discovery of wireless power. Wireless power transmission (WPT) has the potential to be a common source of consumable energy, but it will only receive serious consideration if the transmit and receive systems are extremely efficient and capable of delivering usable amounts of power. Research has been conducted to improve the efficiency and performance of nearly every aspect of WPT systems, but the relatively new field of metamaterials (MTMs) has yet to play a dominate role in improving system performance. A gradient index (GRIN) MTM lens was designed using Ansoft's High Frequency Structure Simulator (HFSS) to improve antenna gain and thereby increase WPT system performance. A simple WPT demonstration system using microstrip patch antennas (MPAs) confirmed the benefits of the GRIN MTM lens. The WPT demonstration system, MPAs, and GRIN MTM lens were constructed and experimentally tested near 2.45 GHz. The theoretical and experimental gain improvement of the MPA due to the GRIN MTM lens is 5.91 dB and 7.06 dB, respectively.

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1 Introduction

1.1 Project Overview

This project demonstrates that metamaterials (MTMs) can improve antenna gain and thereby increase microwave wireless power transmission (WPT) system performance. A MTM lens and simple WPT demonstration system were designed herein to operate at 2.45 GHz with an effective isotropic radiated power (EIRP) above 36 dBm. This frequency was chosen because it's in the ISM band and an FCC license isn't required to transmit EIRP above 36dBm, as defined in Part 15 of Title 47 [1]. WPT system efficiencies greater than 90% have been achieved at 2.45 GHz [2].

A block diagram of the WPT demonstration system is located in Fig. 1.1. The transmitter consists of a voltage controlled oscillator (VCO), voltage controlled attenuator (VCA), band pass filter (BPF), power amplifier, microstrip patch antenna (MPA) and MTM lens. The receiver consists of an MPA, microwave rectifier, low pass filter (LPF), and load.

Fig. 1.1 WPT demonstration system block diagram

The remainder of Chapter 1 discusses the different types of WPT. Chapter 2 presents the MPA theory and design. Chapter 3 introduces MTMs, outlines S-SRR theory, and shows the MTM lens design. Chapter 4 illustrates the WPT demonstration system design. Chapter 5 shows the test results of the MPA, MTM enhanced MPA, and WPT demonstration system. Chapter 6 concludes the project and discusses future work.

1.2 Wireless Power Transmission Methods

Wireless power transmission (WPT) is the transmission of energy between two points without wires. Nikola Tesla first demonstrated WPT in 1891 by means of electrostatic induction using a high-tension induction coil [3]. Three types of WPT methods exist: (1) electromagnetic induction, (2) electrical conduction, and (3) electromagnetic radiation.

Electromagnetic induction WPT is a near field transmission method comprised of two categories, electrodynamic induction and electrostatic induction. Electrodynamic induction transfers energy using mutual inductive coupling between a primary coil and a secondary coil, the same as a power distribution transformer. Electrostatic induction uses alternating current (AC) and capacitive coupling to transfer energy through dielectric material. The efficiency of both forms of electromagnetic induction significantly degrades with distance, therefore limiting their use to close range WPT systems [4].

Electrical conduction WPT is possible using the distributed charge of ground and air method. In this instance, two non-wire conductors transfer energy.

The Earth is one conductor and a capacitive or direct ionized path is the other conductor [5]. This method was the main focus for many of Tesla's experiments and the theoretical basis of the Tesla coil and the Wardencllyffe Tower, shown in Fig. 1.2. This method is theoretically possible because the net resistance of Earth antipodes is less than 1Ω [6]. Oceans and metallic ore conduct energy in this instance. The distributed charge of ground and air method is much more efficient for long distance WPT systems than electromagnetic induction due to the low loss conductive path.

Fig. 1.2 (a) Nikola Tesla sitting next to a Tesla coil [7] (b) the 187-foot Wardencllyffe Tower [2]

WPT using electromagnetic radiation is a far field method typically used for long distance power transmission. The two forms of electromagnetic radiation used in WPT are laser beam transmission and microwave transmission. Laser beam WPT uses a laser transmitter and a solar cell receiver to beam power between two points. Laser energy has a narrow beam cross-section which allows the energy to travel long distances, however converting laser light to electrical energy is inefficient. The atmosphere, clouds, rain, and fog absorb and scatter

laser energy, leading to more inefficiency. Microwave WPT uses antennas to transmit radio waves between two points and microwave rectifiers convert the received energy. Microwave WPT is less prone to absorption and scattering than laser WPT, but antenna gain limits efficiency.

2   Microstrip Patch Antenna

Microstrip patch antennas (MPAs) are the structures that radiate energy from the wireless power transmission demonstration system into the metamaterial lens and capture transmitted energy for use in the microwave rectifier. MPAs are a low cost solution that is easy to manufacture. Despite poor efficiency, low power handling capability, and narrow bandwidth, MPAs provide a simple method to demonstrate the benefits of the metamaterial lens.

2.1   Microstrip Patch Antenna Characteristics

MPAs consist of a microstrip transmission-line strip (or patch) above a ground plane. MPAs are typically fabricated on a dielectric substrate with the radiating patch placed a small fraction of a wavelength away from the ground plane. Dielectric substrates used for microstrip patch antennas are usually in the range of $2.2 \leq \epsilon_r \leq 12$ [8]. Thick substrates with dielectric constants in the lower end of the range are more efficient and have a larger bandwidth because loosely bound fields radiate better into free space. Thin substrates with higher dielectric constants have tightly coupled fields, which is optimal for the feeding network and microwave circuitry since it prevents undesired radiation and coupling. Compromising between the two substrate design options yields optimal balance between antenna and circuit performance.

2.2 Rectangular Microstrip Patch Antenna

The rectangular Microstrip patch antenna is the most common form of MPA, shown in Fig. 2.1. The rectangular MPA is usually operated near resonance to obtain real-valued input impedance [9]. There are four types of feeding methods used to launch energy into MPAs including microstrip line, coaxial probe, aperture coupling, and proximity coupling. Microstrip feed lines, as shown in Fig. 2.1, are easy to fabricate, easy to match, and simple to model. Microstrip feed lines are used in this design due to the aforementioned benefits. Coaxial probe, aperture coupling, and proximity coupling are more difficult to model and therefore omitted from consideration.

2.2.1 Transmission-Line Model

The transmission-line model is the least difficult and least accurate model to analyze rectangular MPAs, but it offers some insight into the radiating mechanism and provides design equations. The design is optimized using a full-wave Finite Element Method (FEM) model in section 2.2.3. The transmission-line model represents the MPA using two slots separated by a half-wavelength, low impedance transmission-line that is open-circuited at the end, shown in Fig. 2.1c.

A standing wave mode forms in the dielectric, shown in Fig. 2.1b, which causes the electric field to be perpendicular to the conductors. The electric field between the two slots in the center of the patch cancel due to opposing polarity, but the fringing electric field outside of the two slots constructively add to

produce a resulting electric field along the x -axis. The resulting E-field causes broadside radiation in the $+z$ -direction.

(a) Geometry for analyzing the edge-fed microstrip patch antenna.

(b) Side view showing the electric fields.

(c) Top view showing the fringing electric fields that are responsible for radiation. The equivalent magnetic surface $\mathbf{M}_s$ currents are also shown.

Fig. 2.1 Half-wavelength rectangular microstrip patch antenna [9]

Some of the fringing electric field is coupled through air and the rest is coupled through the dielectric substrate, which results in a nonhomogeneous field. This causes MPAs, and all microstrip lines, to appear electrically larger. An

effective dielectric constant ϵ_{reff} accounts for fringing and wave propagation. The effective dielectric constant is calculated using the relative dielectric constant ϵ_r of the substrate, height h of the substrate, and width W of the MPA using equation (2.1).

$$\epsilon_{\text{reff}} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + 12 \frac{t}{W} \right)^{-1/2} \quad (2.1)$$

To account for the appearance of being electrically larger, the effective dielectric constant and width-to-height ratio is used to approximate the length extension that fringing causes. Fig. 2.2 shows the physical and effective length extension.

Fig. 2.2 Physical and effective lengths of a rectangular MPA [8]

The normalized extension of length is calculated using

$$\Delta L = 0.412h \frac{(\epsilon_{\text{reff}} + 0.3) \left(\frac{W}{t} + 0.264 \right)}{(\epsilon_{\text{reff}} - 0.258) \left(\frac{W}{t} + 0.8 \right)} \quad (2.2)$$

The effective length caused by fringing fields is therefore

$$L_{eff} = L + 2\Delta L \quad (2.3)$$

Fringing fields and the effective length of the MPA must be considered to ensure that TM_{010} is the dominant resonant mode at the desired resonant frequency. The resonant frequency is calculated for the TM_{010} mode using

$$f_r = \frac{C}{2L_{eff}\sqrt{\epsilon_{reff}}} \quad (2.4)$$

where C is the speed of light in vacuum. A practical patch width that leads to good radiation efficiencies is [8]

$$W = \frac{C}{2f_r} \sqrt{\frac{2}{\epsilon_r + 1}} \quad (2.5)$$

Typical input impedances at the edge of a rectangular resonant patch range from 100Ω to 400Ω [9]. An approximation for the input impedance at the edge of the patch, assuming reactance is zero, is calculated using

$$Z_{in} = 90 \frac{\epsilon_{reff}^2}{\epsilon_{reff} - 1} \left(\frac{L_{eff}}{W} \right)^2 \quad (2.6)$$

As mentioned previously, this design uses a microstrip feed. An impedance transformation matching network is necessary to match the impedance of the MPA with the characteristic impedance of the microstrip feed network. A quarter-wave matching network is used because the MPA input impedance is

ideally assumed as non-reactive. The impedance of the quarter-wave transformer is [10]

$$Z_{qw} = \sqrt{Z_{in} * Z_0} \quad (2.7)$$

2.2.2 Rectangular MPA Design

The dielectric substrate material chosen for the MPAs is FR-4 with a 59mil core and 1oz. copper for a total finished thickness of approximately 62mils. The dimensions of the MPAs are calculated using equations (2.1) through (2.7) with a MATLAB program located in APPENDIX A. The calculated dimensions and selected design parameters of the MPAs are located in Table 2.1. The length and width of the quarter-wave matching network was calculated using LineCalc from Agilent's Advanced Design System (ADS), shown in Fig. 2.3. LineCalc was used to determine the trace width of the 50Ω microstrip feed line characteristic impedance.

Table 2.1 MPA design parameters

f_r (GHz)	2.45
ϵ_r	4.4
ϵ_{reff}	4.1
h (mils)	59
L (mils)	1136
L_{eff} (mils)	1190
ΔL (mils)	27.3
W (mils)	1466
Z_{in} (Ω)	321
Z_{qw} (Ω)	127
W_{qw} (mils)	10.9
L_{qw} (mils)	714
W_{feed} (mils)	112

Component

Type **MLIN** ID **MLIN: MLIN_DEFAULT**

Substrate Parameters

ID **MSUB_DEFAULT**

Er	4.400	N/A
Mur	1.000	N/A
H	59.000	mil
Hu	3.9e+34	mil
T	1.400	mil
Cond	4.1e7	N/A
TanD	0.020	N/A
Rough	0.000	mil
DielectricLossModel	1.000	N/A
FreqForEpsrTanD	1.0e9	N/A
LowFreqForTanD	1.0e3	N/A
HighFreqForTanD	1.0e12	N/A
		N/A

Physical

W **10.901378** mil

L **714.405512** mil

Synthesize **▲** Analyze **▼**

Electrical

Z0 **127.000** Ohm

E_Eff **90.000** deg

Component Parameters

Freq **2.450** GHz

Wall1 mil

Wall2 mil

Fig. 2.3 Agilent ADS quarter-wave matching network dimensions

2.2.3 Full-Wave Model

Ansoft's High Frequency Structure Simulator (HFSS) was used to create a full-wave 3-dimensional model of the MPA designed in the previous section. Full-wave models are the most accurate analysis tool, very versatile, and can compensate for the inaccuracies of the transmission line model. HFSS employs the Finite Element Method (FEM) and an automatic adaptive meshing technique. FEM is a technique used to approximate solutions to boundary problems by dividing the full problem space into smaller regions and represent each sub-region with a local function [11]. The adaptive meshing technique ensures that the mesh

is conformal to the 3-D structure and an appropriate density. All models were simulated using HFSS 14.0.1 on a server running Windows Server 2008 with four Intel Xeon E7-8837 CPUs clocked at 2.67 GHz with 256 GB of RAM and a 3 TB hard drive.

The dimensions from Table 2.1 were used to construct the initial HFSS model on a 6 inch-by-6 inch slab of FR-4, shown in Fig. 2.4. A wave port was used to launch the energy on the microstrip feed line. The wave port dimensions were defined using a “rule of thumb” with the height and width equal to 5-times the substrate height and 5-times the microstrip feed width, respectively [11].

Fig. 2.4 MPA HFSS model

An air box with radiation boundary, or Absorbing Boundary Condition (ABC), surrounds the MPA and dielectric substrate (not shown). The air box is 5 inches tall to make room for the metamaterial lens and ensure a valid solution. Air boxes with ABC boundaries can only absorb energy when the incident angle is

normal or near normal. ABC boundaries must also be at least $\lambda/4$ away from radiating structures to prevent undesired reflections from interfering with the results. Reflections on the ABC boundaries are minimal in this case since the radiated energy is in the broadside direction. A Perfectly Matched Layer (PML) boundary could have been used to minimize reflections and ensure a valid solution, since they use artificial layers to absorb all outgoing energy. A PML wasn't used because they make it more difficult for the iterative solver to reach convergence, therefore requiring more memory and more time.

The solution setup has a *Maximum Delta S* of 0.01 with a minimum of three converged passes. The *Maximum Delta S* is the magnitude of S-parameter change between two consecutive passes. Setting the converged pass minimum to three helps ensure that the *Maximum Delta S* has sufficiently converged. The finite element mesh, located in Fig. 2.5, shows satisfactory convergence as shown by the tetrahedra density near the MPA.

The MPA dimensions were optimized to ensure ideal gain and reflection coefficient performance at the desired resonant frequency. A reduction in MPA length was necessary to center the resonant frequency at 2.45 GHz. The length and width of the quarter-wave transformer had to be increased to improve matching to the 50 Ω microstrip feed characteristic impedance. The resulting optimized dimensions are located in Table 2.2.

Fig. 2.5 HFSS post-convergence MPA mesh plot

Table 2.2 Optimized MPA design parameters

	Calculated	Optimized	% Difference
L (mils)	1136	1101	3.08
W_{qw} (mils)	10.9	41	276
L_{qw} (mils)	714	724	1.40

The resonant frequency of the optimized MPA was shifted 75 MHz higher in comparison to the original design, only 4 MHz away from the design goal of 2.45 GHz, as shown in Fig. 2.6. The maximum return loss of the optimized MPA was increased to 34.6 dB from the original design, as shown in Fig. 2.6, demonstrating an improved match to the characteristic impedance of the system. The peak boresight gain of the optimized MPA was increased 0.74 dB from the

original design, shown in Fig. 2.7. Both Fig. 2.8 and Fig. 2.9 illustrate the improved impedance match of the optimized MPA to 50Ω . Fig. 2.10 shows the original and optimized MPA radiation patterns. The HPBW in the E-plane was increased 14 degrees due to the decrease in patch length. Table 2.3 summarizes the original and optimized MPA performance characteristics.

Table 2.3 HFSS Optimized MPA performance characteristics comparison

	Original	Optimized
f_r (GHz)	2.379	2.454
Min S11 (dB)	-8.73	-43.3
Peak Boresight Gain (dB)	2.16	2.9
MPA Port Z at 2.45 GHz [real] (Ω)	13.2	45.2
MPA Port Z at 2.45 GHz [imag] (Ω)	-40.0	2.29
E-Plane HPBW (deg)	85	99
H-Plane HPBW (deg)	72	74

Fig. 2.6 HFSS optimized MPA input reflection coefficient

Fig. 2.7 HFSS optimized MPA boresight swept gain

Fig. 2.8 HFSS optimized MPA real impedance

Fig. 2.9 HFSS optimized imaginary impedance

Fig. 2.10 HFSS optimized MPA radiation pattern

HFSS has the ability to overlay field and radiation plots on the solution space. Magnitude and vector representations of the electric field, magnetic field, surface currents, and other parameters can be plotted and animated to get an intuitive understanding of field interactions with the structure in the solution space. Fig. 2.11 shows a magnitude overlay of the electric field located on the conductive surface of the MPA, microstrip feed line, and quarter-wave matching network. Fig. 2.12 shows a cross section of the electric field magnitude across the X-Z plane. Fig. 2.13 shows a 3-dimensional view of the electric field magnitude overlaid within the air box surrounding the MPA. Fig. 2.12 and Fig. 2.13 clearly show the E-field radiating into free space.

Fig. 2.11 Electric field magnitude overlay on the MPA conductive surfaces

Fig. 2.12 Electric field magnitude overlay on the X-Z plane

Fig. 2.13 Electric field magnitude overlay in the air box surrounding the MPA

3.1 Introduction to Metamaterials

Metamaterials (MTMs) are defined as artificial effectively homogeneous electromagnetic structures with properties not readily found in nature [12]. A structure is considered effectively homogeneous when the cell size of the metallic inclusions is less than a quarter-wavelength. The effective homogeneity limit of a quarter-wavelength ensures that incident electromagnetic energy will exhibit refraction instead of diffraction and scattering. An effectively homogeneous structure is considered electromagnetically uniform along the direction of wave propagation and is defined by the permittivity ϵ , permeability μ , and the refractive index n by

$$n = \pm\sqrt{\epsilon_r\mu_r} \quad (3.1)$$

where ϵ_r and μ_r are the relative permittivity and permeability related to the free space permittivity and permeability, respectively, by

$$\epsilon_0 = \frac{\epsilon}{\epsilon_r} = 8.854 * 10^{-12} \quad (3.2)$$

$$\mu_0 = \frac{\mu}{\mu_r} = 4\pi * 10^{-7} \quad (3.3)$$

Materials with simultaneously negative permittivity ϵ and permeability μ don't exist naturally and are therefore considered metamaterials. In 1967, Victor Veselago postulated the existence of a material with simultaneously negative permittivity ϵ and permeability μ [13]. Veselago concluded that Maxwell's

equations permit the existence of such material and the electric field vector $\mathbf{E}$, magnetic field vector $\mathbf{H}$, and the wave vector $\mathbf{k}$ form a left-handed triplet [14]. Left-handed material exhibits several unique properties, such as backward wave propagation, reversal of Snell's law, reversal of some boundary conditions, reversal of the Doppler Effect, reversal of Vavilov-Cerenkov radiation, and other phenomena. Even though wave propagation is in the reverse direction, the Poynting vector $\mathbf{S}$ maintains a right-handed relationship with $\mathbf{E}$ and $\mathbf{H}$, causing the phase and group velocities to be antiparallel. Fig. 3.1 shows the four possible sign combinations for ϵ and μ .

Fig. 3.1 Material permittivity-permeability and refractive index diagram [12]

3.1.1 Wave Propagation in Left-Handed Media

Maxwell's wave equation is introduced to illustrate backward wave propagation in left-handed material:

$$\left(\nabla^2 - \frac{n^2}{C^2} \frac{\partial^2}{\partial t^2} \right) \psi = 0 \quad (3.4)$$

where n is the refractive index, C is the speed of light in vacuum, and $n^2/C^2 = \epsilon\mu$ [15]. The index of refraction is unaffected by simultaneous sign change of the permittivity and permeability and therefore low-loss left-handed media is transparent to electromagnetic energy. Even though left-handed media appears transparent and equation (3.4) is unchanged in double-negative media, the differential form of Faraday's and Ampère's laws, as shown in equations (3.5) and (3.6), are affected.

$$\nabla \times \mathbf{E} = -j\omega\mu\mathbf{H} \quad (3.5)$$

$$\nabla \times \mathbf{H} = j\omega\epsilon\mathbf{E} \quad (3.6)$$

For plane-wave electric and magnetic fields represented by

$$\mathbf{E} = \mathbf{E}_0 e^{-j\mathbf{k}\cdot\mathbf{r} + j\omega t} \quad (3.7)$$

$$\mathbf{H} = \mathbf{H}_0 e^{-j\mathbf{k}\cdot\mathbf{r} + j\omega t} \quad (3.8)$$

equations (3.5) and (3.6) reduce to the following

$$\mathbf{k} \times \mathbf{E} = \omega\mu\mathbf{H} \quad (3.9)$$

$$\mathbf{k} \times \mathbf{H} = -\omega\epsilon\mathbf{E} \quad (3.10)$$

Therefore, positive values of ϵ and μ form a right-handed triplet between $\mathbf{E}$, $\mathbf{H}$, and $\mathbf{k}$. Similarly, negative values of ϵ and μ form a left-handed triplet.

Even though wave propagation is in the backward direction, the direction of real-valued, time-averaged power propagation, the Poynting vector, is in the forward direction as shown by

$$\mathbf{S} = \frac{1}{2} \text{Re}(\mathbf{E} \times \mathbf{H}^*) \quad (3.11)$$

3.1.2 Boundary Conditions

The boundary conditions which describe the interface between two media are derived from Maxwell's equations and shown in general form as [12]

$$\mathbf{n} \cdot (\mathbf{D}_2 - \mathbf{D}_1) = \rho_{se} \quad (3.12)$$

$$\mathbf{n} \cdot (\mathbf{B}_2 - \mathbf{B}_1) = \rho_{sm} \quad (3.13)$$

$$\mathbf{n} \times (\mathbf{E}_2 - \mathbf{E}_1) = -\mathbf{M}_s \quad (3.14)$$

$$\mathbf{n} \times (\mathbf{H}_2 - \mathbf{H}_1) = \mathbf{J}_s \quad (3.15)$$

where ρ_{se} is the electric surface charge density on the interface, ρ_{sm} is the fictitious magnetic surface charge density on the interface, and $\mathbf{n}$ is the unit vector normal to the interface pointing from medium 1 to medium 2. In the absence of charges at the interface ($\rho_{se} = \rho_{sm} = 0$) the normal components of $\mathbf{D}$ and $\mathbf{B}$ are continuous, shown in equations (3.12) and (3.13). In the absence of sources at the interface ($\mathbf{J}_s = \mathbf{M}_s = 0$) the normal components of $\mathbf{D}$ and $\mathbf{B}$ are continuous, shown in equations (3.14) and (3.15). Applying the aforementioned conditions reduces equations (3.12) through (3.15) to the following equations

$$D_{1n} = D_{2n} \quad (3.16)$$

$$B_{1n} = B_{2n} \quad (3.17)$$

$$E_{1t} = E_{2t} \quad (3.18)$$

$$H_{1t} = H_{2t} \quad (3.19)$$

where the indexes n and t correspond to normal and tangential components.

The electric displacement field $\mathbf{D}$ and the magnetic flux density $\mathbf{B}$ are directly affected by sign changes of the permittivity and permeability, therefore equations (3.16) and (3.17) show that the interface between right and left-handed media have boundary conditions with antiparallel normal components. $\mathbf{E}$ and $\mathbf{H}$ are unaffected by sign changes for ϵ and μ , therefore equations (3.18) and (3.19) show the boundary conditions for tangential components are parallel. Fig. 3.2 shows the special case of an interface between right-handed and left-handed media.

Fig. 3.2 Boundary conditions at the interface between right-handed and left-handed media [12]

3.1.3 Negative Refraction and Snell's Law

The negative refractive index of left-handed material causes a reversal in Snell's law of refraction. To illustrate the effects of negative refraction, consider the cases where two right-handed media are interfaced and right-handed and left-handed media are interfaced, as shown in Fig. 3.3a and Fig. 3.3b, respectively.

Electromagnetic energy

Fig. 3.3 Refraction of an electromagnetic wave upon RH-RH and RH-LH interfaces [12]

incident upon an interface between two media of the same handedness will experience positive refraction and a positive refraction angle. Electromagnetic energy incident upon two media of different handedness will experience negative refraction and a negative refraction angle. This result is clear when considering all four sign cases for Snell's law of refraction, shown in equation (3.20).

$$n_1 \sin \theta_1 = n_2 \sin \theta_2 \quad (3.20)$$

For the case when both materials at the interface are of the same handedness, whether right or left, Snell's law maintains conventional positive refraction due to canceling of negative signs, if present. When the two materials are of opposite handedness, negative signs don't cancel and Snell's law describes negative refraction.

3.2 S-Shaped Split Ring Resonator Theory

The S-Shaped Split Ring Resonator (S-SRR) is a metamaterial structure discovered by Chen et al. in 2004 [16]. S-SRRs are unique from the metamaterial structures discovered prior, and are composed of only an S-shaped structure and an inverted image. Previously realized metamaterial structures were composed of two structure types, split ring resonators (SRRs) and rod arrays [17], [18]. S-SRRs can achieve simultaneously negative permittivity and permeability within the same frequency band and can therefore exhibit left-handed behavior. Fig. 3.4a shows an array of S-SRRs and Fig. 3.4b shows a unit cell consisting of the S-shaped structure with its image [19].

The S-SRR unit cell has a dimensions of a in the x -direction, b in the z -direction, and l in the y -direction. The area each unit cell occupies is therefore, $S = ab$. The unit cell, which appears as a figure-eight when viewed from the x - z plane, is divided into three regions, as shown in Fig. 3.4b, labeled F_1 , F_2 , and F_3 . F_1 is the fractional volume occupied by the top loop (Loop 1) in the figure-eight pattern, F_2 is the fractional volume occupied by the bottom loop (Loop 2) in the

figure-eight pattern, and F_3 is the fractional volume surrounding both loops.

Equation (3.21) relates the fractional volumes to the total volume.

Fig. 3.4 S-Shaped Split Ring Resonator structure [19]

$$F_1 + F_2 + F_3 = 1 \quad (3.21)$$

A time-varying magnetic field H_0 applied in the y-direction causes current to flow in the two loops of the structure, denoted by I_1 and I_2 . The total applied magnetic field, fractional volumes of the structure, and the electric current density relate to the magnetic fields in each region by

$$H_1 - H_2 = J_1 - J_2 \quad (3.22)$$

$$H_1 - H_3 = J_1 \quad (3.23)$$

$$H_1 F_1 + H_2 F_2 + H_3 F_3 = H_0 \quad (3.24)$$

where the units of the currents are Ampere's per unit length in the y-direction

$$J_1 = \frac{I_1}{l} \quad (3.25)$$

$$J_2 = \frac{I_2}{l} \quad (3.26)$$

The magnetic fields within each fractional volume are found using equations

(3.22) through (3.24) and are shown as

$$H_1 = H_0 + (1 - F_1)J_1 - F_2J_2 \quad (3.27)$$

$$H_2 = H_0 - F_1J_1 + (1 - F_2)J_2 \quad (3.28)$$

$$H_3 = H_0 - F_1J_1 - F_2J_2 \quad (3.29)$$

The differential form of Faraday's law relates the electromotive force (emf) to the fractional volume and magnetic field within by

$$\text{Loop 1: } emf_1 = -\frac{\partial}{\partial t}(\mu_0 H_1 F_1 S) \quad (3.30)$$

$$= \sigma_s I_1 + \frac{1}{C_s} \int I_1 dt + \frac{1}{C_m} \int (I_1 + I_2) dt$$

$$\text{Loop 2: } emf_2 = -\frac{\partial}{\partial t}(\mu_0 H_2 F_2 S) \quad (3.31)$$

$$= \sigma_s I_2 + \frac{1}{C_s} \int I_2 dt + \frac{1}{C_m} \int (I_1 + I_2) dt$$

where σ_s is the resistance of the metal composing each loop, C_m is the capacitance between the metal strips in the middle of the structure, and C_s is the capacitance between the metal strips on the top and bottom of the structure. Equation (3.32)

relates the capacitance between the metal strips on the top, middle, and bottom of the structure to the S-SRR dimensions by

$$C_s = C_m = \varepsilon_0 \frac{hc}{d} + \varepsilon_0 \frac{hc}{l-d} \quad (3.32)$$

The currents in equations (3.30) and (3.31) are then represented using equations (3.27) through (3.29). Substituting $\partial/\partial t$ with $-j\omega$ and $\int dt$ with $1/-j\omega$, the result is

$$j\omega\mu_0 H_1 F_1 S - \sigma_s J_1 l + \frac{J_1 l}{j\omega C_s} + \frac{(J_1 + J_2)l}{j\omega C_m} = 0 \quad (3.33)$$

$$j\omega\mu_0 H_2 F_2 S - \sigma_s J_2 l + \frac{J_2 l}{j\omega C_s} + \frac{(J_1 + J_2)l}{j\omega C_m} = 0 \quad (3.34)$$

The effective permeability is then calculated using equations (3.25), (3.26), (3.33), and (3.34) as shown

$$\begin{aligned} \mu_{eff} \\ = 1 \\ - \frac{(\omega\mu_0 S)^2 F_2 F_1 (F_1 + F_2) - \mu_0 S \left[\frac{(F_1^2 + F_2^2)}{C_s} + \frac{l(F_1 - F_2)^2}{C_m} \right] + jA(\sigma)}{(\omega\mu_0 S)^2 F_1 F_2 - \mu_0 S (F_1 + F_2) \left(\frac{l}{C_s} + \frac{l}{C_m} \right) + \frac{l}{\omega^2 C_s} \left(\frac{l}{C_s} + \frac{2l}{C_m} \right) - B(\sigma) + jC(\sigma)} \end{aligned} \quad (3.35)$$

where

$$A(\sigma) = \omega\mu_0 S (F_1^2 + F_2^2) \sigma_s l \quad (3.36)$$

$$B(\sigma) = (\sigma_s l)^2 \quad (3.37)$$

$$C(\sigma) = \left[\omega\mu_0 S (F_1 + F_2) - \frac{2}{\omega} \left(\frac{l}{C_s} + \frac{l}{C_m} \right) \right] \sigma_s l \quad (3.38)$$

The special case where the S-SRR is symmetric and the fraction volumes F_1 and F_2 are the same ($F_1 = F_2 = F$), the effective permeability reduces to

$$\mu_{eff} = 1 - \frac{2F + jD(\sigma)}{1 - \frac{1}{\omega^2 \mu_0 FS} \left(\frac{l}{C_s} + \frac{2l}{C_m} \right) - E(\sigma) + jG(\sigma)} \quad (3.39)$$

where

$$X = (\omega \mu_0 FS)^2 \left(1 - \frac{l}{\omega \mu_0 FS C_s} \right) \quad (3.40)$$

$$D(\sigma) = \frac{A(\sigma)}{X} \quad (3.41)$$

$$E(\sigma) = \frac{B(\sigma)}{X} \quad (3.42)$$

$$G(\sigma) = \frac{C(\sigma)}{X} \quad (3.43)$$

Unlike the case when fractional volumes F_1 and F_2 are different, a symmetric S-SRR has singular magnetic resonant and magnetic plasma frequencies. The magnetic resonant frequency is calculated using

$$\omega_{m0} = \sqrt{\frac{1}{\mu_0 FS} \left(\frac{l}{C_s} + \frac{2l}{C_m} \right)} \quad (3.44)$$

The magnetic plasma frequency is calculated using

$$\begin{aligned}\omega_{mp} &= \sqrt{\frac{1}{\mu_0 FS(1-2F)} \left(\frac{l}{C_s} + \frac{2l}{C_m} \right)} \\ &= \omega_{m0} \sqrt{\frac{1}{1-2F}}\end{aligned}\tag{3.45}$$

3.3 S-Shaped Split Ring Resonator Array Design

The design equations from the previous section were used to create a MATLAB program to calculate the expected behavior of an S-SRR, located in APPENDIX B. Dimensional ratios set forth by Chen et al. in [19] were used as a basis for the S-SRR dimensions in this design. All dimensions were defined and scaled as a function of the unit cell height a , except for the distance d between the mirrored S-structures and the length l between the S-SRRs. The distance d was chosen as 59 mils, the thickness of the FR-4 dielectric substrate that the S-SRRs was fabricated on. The length l was chosen as 500 mils to reduce the number of strips in the final S-SRR array. The cell height a was swept until the magnetic resonant frequency occurred near, but slightly less than, 2.45 GHz so that the permeability would be zero at the desired frequency of operation. Table 3.1 lists the theoretical S-SRR dimensions that achieve a zero-valued permeability near 2.45 GHz. Fig. 3.5 shows the theoretical permeability frequency response.

Table 3.1 Theoretical S-SRR unit cell dimensions and frequency characteristics

a (mils)	1400
b (mils)	700
w (mils)	1050
h (mils)	560
c (mils)	70
d (mils)	59
l (mils)	500
f_{m0} (GHz)	2.45
f_{mp} (GHz)	3.87

Fig. 3.5 S-SRR unit cell theoretical effective permeability

A 3-dimensional full-wave model of a unit cell in was created in HFSS using the dimensions listed in Table 3.1, shown in Fig. 3.6. The S-SRR was modeled using a loaded waveguide approach, where the boundaries of the unit cell along the Y-Z plane are Perfect-E and the boundaries along the X-Z plane are Perfect-H, shown highlighted in blue in Fig. 3.7a and Fig. 3.7b. These boundary conditions ensure that a perfect plane wave excites the structure uniformly. The unit cell boundaries along the X-Y plane were defined as wave ports, shown highlighted in red in Fig. 3.7c.

Fig. 3.6 S-SRR HFSS model of unit cell on 59 mil FR-4

Fig. 3.7 S-SRR unit cell HFSS boundaries and excitations (a) Perfect-E (b) Perfect-H (c) Wave Ports

The S-SRR unit cell S-parameters were modeled using a loaded waveguide configuration. Ziolkowski's modified Nicholson-Ross-Weir method extracts the effective permittivity and permeability using the S-parameters [20], [21], [22]. A parameter extraction MATLAB program, listed in APPENDIX C, was written to extract and display the permittivity, permeability, refractive index, intrinsic impedance, and wave reflection coefficient.

The presence of FR-4 substrate between the mirrored S-shaped structures significantly alters the magnetic permeability response of the S-SRR unit cell. The original dimensions listed in Table 3.1 were optimized so the permeability response from the parameter extraction program matched the theoretical permeability response in Fig. 3.5. The optimized S-SRR unit cell dimensions are listed in Table 3.2 and the extracted parameters are shown in Fig. 3.8 through Fig. 3.11.

Table 3.2 S-SRR optimized unit cell dimensions

a (mils)	1050
b (mils)	525
w (mils)	787.5
h (mils)	420
c (mils)	52.5
d (mils)	59
l (mils)	500
f_{m0} (GHz)	2.45
f_{mp} (GHz)	3.87

The S-SRR unit cell was optimized so the permittivity and permeability resonated in unison and the real parts of both were approximately zero at the desired operating frequency of 2.45 GHz, shown in Fig. 3.8. Equation (3.1) shows that the real part of the refractive index will also be zero, as seen in Fig. 3.9. Since both the permittivity and permeability resonate in unison, the normalized wave impedance will approach unity as defined by equation (3.46) for a lossless medium. As the intrinsic impedance of the material approaches unity, the reflection coefficient of the material will approach zero, as seen in Fig. 3.11.

$$\eta = \sqrt{\frac{\mu}{\epsilon}} \quad (3.46)$$

Fig. 3.8 Extracted permittivity and permeability of the optimized S-SRR unit cell with $a = 1050\text{mils}$

Fig. 3.9 Extracted refractive index of the optimized S-SRR unit cell with $a = 1050\text{mils}$

Fig. 3.10 Extracted normalized wave impedance of the optimized S-SRR unit cell with $a = 1050\text{mils}$

Fig. 3.11 Extracted material reflection coefficient of optimized S-SRR unit cell with $a = 1050\text{mils}$

Zero refractive index metamaterials have the potential to significantly increase the gain of the MPA designed in chapter 2 [23], [24]. Solving Snell's law for electromagnetic energy originating from within zero refractive index metamaterials shows that all waves exiting the structure will have an angle of refraction at or near zero, regardless of the incident angle. This concept, as illustrated in Fig. 3.12, can effectively transform an incident spherical wave into a plane wave.

Fig. 3.12 Wave propagation from energy inside metamaterial with zero refractive index [23]

The S-SRR unit cell dimensions from Table 3.2 were used as a basis for the S-SRR array shown in Fig. 3.13. The HFSS model, shown in Fig. 3.14, was created using the loaded waveguide approach, similar to how the unit cell was analyzed, with perfect-E boundaries and perfect-H boundaries on the Y-Z and X-Z planes, respectively. Wave ports were assigned on both sides of the array in the X-Y plane, similar to the unit cell. Unfortunately, the permittivity and permeability of the S-SRR array resonate differently than the unit cell and the ideal response shown in Fig. 3.8 through Fig. 3.11 doesn't occur. The extracted permittivity and permeability of the array are shown in Fig. 3.15.

Fig. 3.13 S-SRR Array in HFSS

Fig. 3.14 S-SRR array HFSS boundaries (a) Perfect-E (b) Perfect-H

Optimizing the S-SRR array is much more troublesome than the unit cell. Typical simulation time for each parametric analysis takes approximately 150 hours (CPU time) and has a peak memory usage of 180 GB of ram. This can be attributed to the large wave ports and their proximity to the resonating elements.

Fig. 3.15 shows a dual band condition with electric and magnetic resonances near 2.33 GHz and 2.59 GHz. Several iterations were attempted to realign the electric and magnetic resonant frequencies to 2.45 GHz, but the design never converged. Two such instances are shown in Fig. 3.16 and Fig. 3.17. Increasing the size of each unit cell in the array causes the resonant frequencies to decrease, but the permittivity and permeability of the upper resonant band no longer resonate in unison and begin to diverge with increasing size of each unit cell, shown in Fig. 3.16. Decreasing the size of each unit cell in the array causes the resonant frequencies to increase, but the permittivity and permeability of the lower resonant band no longer resonate negative, shown in Fig. 3.17.

Fig. 3.15 Extracted permittivity and permeability of the S-SRR array with $a = 1050\text{mils}$

Fig. 3.16 Extracted permittivity and permeability of the S-SRR array with $a = 1100\text{mils}$

Fig. 3.17 Extracted permittivity and permeability of the S-SRR array with $a = 975\text{mils}$

3.4 MTM Enhanced MPA

As mentioned in the previous section, iterating the S-SRR array design is time consuming when using the loaded waveguide approach. Rather than spending weeks waiting for convergence that might never occur, the S-SRR array performance was simulated using the intended operating configuration.

Simulating the S-SRR array in conjunction with the rectangular microstrip patch antenna reduces the typical simulation time of the model from 150 hours (CPU time) down to 16 hours (CPU time).

3.4.1 Metamaterial Enhanced MPA Optimization

The metamaterial enhanced MPA HFSS model is shown in Fig. 3.18. To optimize MPA boresight gain, the S-SRR array location was parameterized and swept. The array location was swept from 0.25 inches to 4 inches away from the microstrip patch. Optimized gain occurs when the S-SRR array is 2.75 inches (0.571λ) away from the MPA. The radiated wave from the MPA has difficulty forming when the array is closer than 2.75 inches. The lens captures less energy when the array is farther than 2.75 inches due to spherical spreading.

After the optimal array location was found, the size of the array was swept to find the dimensions which achieve maximum boresight gain at the desired operating frequency. As mentioned previously, all dimensions of the array were defined as functions of the cell height a , so the parametric optimization was a function of only one variable. A select few iterations of the boresight swept gain optimization are located in Fig. 3.19 through Fig. 3.22. A peak realized gain of

8.81dB near 2.45 GHz occurs when the unit cells in the S-SRR array have the dimensions listed in Table 3.3. The S-SRR lens effectively increases the simulated MPA realized peak gain of 2.90 dB by 5.91 dB. Table 3.4 shows a summary of performance characteristics for the MPA and MTM enhanced MPA. An overlay of the radiation pattern on the MTM enhanced MPA model is located in Fig. 3.23.

Fig. 3.18 Metamaterial enhanced MPA HFSS model

Fig. 3.19 MTM enhanced MPA HFSS maximum boresight gain optimization

Fig. 3.20 MTM enhanced MPA input reflection coefficient

Fig. 3.21 MTM enhance MPA input impedance (real)

Fig. 3.22 MTM enhanced MPA input impedance (imaginary)

Table 3.3 S-SRR unit cell dimensions for MTM enhanced MPA maximum swept gain optimization

a (mils)	890
b (mils)	500
w (mils)	667.5
h (mils)	356
c (mils)	44.5
d (mils)	59
l (mils)	500
MTM location (in.)	2.75

Fig. 3.23 MTM enhanced MPA radiation pattern overlay

3.4.2 MTM Enhanced MPA Lens Operation

Electric and magnetic field overlays are shown in Fig. 3.24 through Fig. 3.28. When the peak electric field created by the MPA excites the S-SRR array, shown in Fig. 3.24 and Fig. 3.25, the S-SRR unit cells become polarized along the x -axis and an electric field forms between adjacent unit cells. When the peak magnetic field created by the MPA excites the S-SRR array, shown in Fig. 3.26, current flows in the S-SRR metallic strips and the magnetic field is concentrated inside the two loops of each unit cell. The MTM focusing ability on the E-field is shown in Fig. 3.27 and Fig. 3.28. Similar focusing of the H-field occurs, but the figures are omitted.

Fig. 3.24 MTM enhanced MPA E-field magnitude on X-Z plane, 0-degree phase shift

Fig. 3.25 MTM enhanced MPA E-field magnitude on X-Z plane cut, 0-degree phase shift

Fig. 3.26 MTM enhanced MPA H-field magnitude on X-Z plane, 0-degree phase shift

Fig. 3.27 MTM enhanced MPA E-field magnitude on X-Z plane, 90-degree phase shift

Fig. 3.28 MTM enhanced MPA E-field magnitude on Y-Z plane, 90-degree phase shift

The previous figures clearly show that the MTM array is resonating with applied electromagnetic fields at 2.45 GHz and that the applied waves are effectively focused. The loaded waveguide S-SRR array model was revisited using parameters located in Table 3.3 to understand the operating characteristics of the case with the highest boresight swept gain. The extracted permittivity, permeability, refractive index, intrinsic impedance, and reflection coefficient are shown in Fig. 3.29 through Fig. 3.32. The extracted parameters show that the S-SRR array characteristics are nowhere near left-handed operation. The permittivity and permeability don't resonate in unison which causes the normalized intrinsic impedance to be different than free space, resulting in a non-zero reflection coefficient.

When analyzing the extracted material characteristics, the MTM lens apparent shortcomings make it seem like the lens wouldn't contribute to focusing and would reflect incident energy. The Poynting vector overlay in HFSS was used to illustrate which portions of the S-SRR array contributed to focusing, shown in Fig. 3.33 and Fig. 3.34. The overlay shows that the center strips in the S-SRRs are the main contributors to the focusing ability of the lens and that the remaining pieces of the S-SRRs either scatter incident energy or attenuate it.

Fig. 3.29 Extracted permittivity and permeability of S-SRR array with dimension in Table 3.3

Fig. 3.30 Extracted refractive index of S-SRR array with dimension in Table 3.3

Fig. 3.31 Extracted Wave Impedance of S-SRR array with dimension in Table 3.3

Fig. 3.32 Extracted reflection coefficient of S-SRR array with dimension in Table 3.3

An important characteristic of the MTM lens revealed by the Poynting vector overlay is that the maximum power density is directly above the MPA and radially decays from the z -axis. The Poynting vectors in the center strips of each S-SRR are all approximately normal to the surface of the effectively homogeneous lens. Power flow that is normal to the surface of a lens which decays radially from the center is characteristic of a gradient refractive index (GRIN) lens.

Fig. 3.33 MTM enhanced MPA Poynting vector in MTM, isometric view

Fig. 3.34 MTM enhanced MPA Poynting vector in MTM, top view

Planar GRIN lenses are well known to transform spherical waves into plane waves [25], [26], [27]. The S-SRR array has a gradient refractive index because the refractive index of each unit cell changes as the angle of incident energy increases. To verify the GRIN mode of operation, the S-SRR unit cell HFSS model shown Fig. 3.6 was rotated along the x -axis and the refractive index was extracted. As shown in Fig. 3.35, the refractive index decreases as the incident angle increases.

Fig. 3.35 Extracted refractive index of S-SRR unit cell vs. incident angle for $a = 1050$

3.4.3 MTM Enhanced MPA and Optimized MPA Comparison

The performance comparison between the MPA and MTM enhanced MPA is summarized in Table 3.4 and shown in Fig. 3.36 through Fig. 3.40. The peak boresight gain is increased by 5.91 dB and the resonant frequency shifts 9 MHz. The E-plane and H-plane half-power beam widths are significantly reduced with the MTM lens by 57 degrees and 28 degrees, respectively. The MTM lens reduces the match of the MPA to the $50\ \Omega$ characteristic impedance of the system as seen in the change in port impedance and decreased return loss.

Table 3.4 MPA and MTM enhanced MPA performance comparison

	MPA	MTM MPA
f_r (GHz)	2.454	2.445
Min S11 (dB)	-43.3	-18.8
Peak Boresight Gain (dB)	2.90	8.81
MPA Port Z at 2.45 GHz [real] (Ω)	45.2	42.4
MPA Port Z at 2.45 GHz [imag] (Ω)	2.29	-8.23
E-Plane HPBW (deg)	99	42
H-Plane HPBW (deg)	72	44

Fig. 3.36 MTM enhanced MPA and optimized MPA swept gain comparison

Fig. 3.37 MTM enhanced MPA and optimized MPA radiation pattern comparison

Fig. 3.38 MTM enhanced MPA and optimized MPA input reflection coefficient comparison

Fig. 3.39 MTM enhanced MPA and optimized MPA input impedance (real) comparison

Fig. 3.40 MTM enhanced MPA and optimized MPA input impedance (imaginary) comparison

4.1   WPT Demonstration System Design Considerations

A basic demonstration system to transmit and receive wireless power was constructed to verify the metamaterial lens benefits. Efficiency is typically a main concern of WPT systems; however, design considerations for this system did not include efficiency. The power added efficiency (PAE) of the amplifier, the spectral purity of the oscillator, and the minimum insertion loss of the band pass filter and voltage controlled attenuator were therefore not driving factors in the design decisions. The main goal was to transmit enough power in the desired frequency band to demonstrate the MTM lens gain improvement. The system was designed to transmit approximately 36dBm EIRP at 2.45 GHz using inexpensive commercial off the shelf (COTS) components.

A four-layer PCB stack-up was chosen with 14 mil FR-4 cores/prepreg and 1 oz. copper for an approximate finished thickness of 62 mils, shown in Fig. 4.1. A coplanar waveguide transmission line was chosen for the 50 Ω RF path. The width of the transmission line (25.6 mils) and the spacing to the coplanar ground plane (22 mils) was determined using Agilent ADS. Coplanar waveguide has superior noise rejection and isolation in comparison to microstrip transmission line.

Fig. 4.1 WPT demonstration system PCB stack-up

4.2 Voltage Controlled Oscillator

A Mini-Circuits ROS-2625-119+ voltage controlled oscillator (VCO) was chosen as the signal source for the transmitter, shown in Fig. 4.2. This VCO was chosen primarily for its frequency range, tuning voltage sensitivity, and output power. The frequency range of this VCO will allow for full characterization of the system and the MTM enhanced MPA. The ROS-2625-119+ is packaged in a dielectric substrate carrier covered with a metal RF shield.

Fig. 4.2 Mini-Circuits VCO ROS-2625-119+ [28]

The EAGLE CAD schematic and layout for the ROS-2625-119+ design is located in Fig. 4.3 and Fig. 4.4, respectively. Bypass capacitors are located on the supply line and tuning voltage line to ensure RF doesn't couple to these traces unintentionally. The trace width of the VCO control voltage signal is 32 mils.

Fig. 4.3 Mini-Circuits VCO ROS-2625-119+ EAGLE CAD schematic

Fig. 4.4 Mini-Circuits VCO ROS2625-119 EAGLE CAD layout

4.3 Voltage Controlled Attenuator

An RFMD RFSA2013 voltage controlled attenuator (VCA), shown in Fig. 4.5, was included in the WPT demonstration system for two reasons: (1) to ensure the input power level into the amplifier didn't exceed the specifications or cause it to go into saturation and (2) to simulate path loss in the WPT demonstration system without moving the transmitter and receiver further apart. The RFSA2013 is an analog VCA with an attenuation range of 33 dB and a minimum insertion loss of 2.6 dB. It is offered in a small 3mm by 3mm square QFN package with a ground paddle.

Fig. 4.5 RFMD RFSA2013 (a) QFN package close-up (b) functional block diagram [29]

The RFSA2013 EAGLE CAD schematic is shown in Fig. 4.6. DC blocking capacitors C201 and C202 are located on the RF input and output nets to prevent DC voltage from the VCA getting into other components in the RF path. Bypass capacitors and a DC biasing resistor are shown in the top portion of Fig. 4.6. The mode signal (pin 16) controls the attenuation slope of the device. A logic

high input ($V_{\text{MODE}} = V_{\text{DD}}$) results in a positive attenuation slope, where decreasing control voltage results in greater attenuation. A logic low input ($V_{\text{MODE}} = \text{GND}$) results in a negative attenuation slope, where increasing control voltage results in greater attenuation. A positive attenuation slope was chosen for this design.

Expected transmission coefficient characteristics are shown in Fig. 4.7.

The RFSA2013 EAGLE CAD layout is shown in Fig. 4.8. The RF transmission line characteristics are the same dimensions as the $50\ \Omega$ coplanar waveguide transmission line used in the layout for the VCO. The traces for the V_{DD} and mode pins are 10 mils wide. The attenuation control voltage trace width is 32 mils.

Fig. 4.6 RFMD voltage controlled attenuator RFSA2013 EAGLE CAD schematic

Fig. 4.7 RFMD RFSA2013 published insertion loss vs. frequency with positive attenuation slope [29]

Fig. 4.8 RFMD voltage controlled attenuator RFSA2013 EAGLE CAD layout

4.4 Ceramic Band Pass Filter

A ceramic band pass filter (BPF) from Johanson Technologies, Inc. was chosen to remove harmonic signals generated from the VCO. The JTI 2450BP39D100C was chosen because of its small size, low cost, and relatively low insertion loss. Ceramic BPFs use quarter-wave resonators embedded in ceramic with a capacitive coupling network [30]. Expected transmission and reflection characteristics are shown in Fig 4.9. The EAGLE CAD schematic and layout are shown in Fig. 4.10 and Fig. 4.11, respectively.

Fig. 4.9 JTI ceramic band pass filter 2450BP39D100C published performance [31]

Fig. 4.10 JTI ceramic band pass filter 2450BP39D100C EAGLE CAD schematic

Fig. 4.11 JTI ceramic band pass filter 2450BP39D100C EAGLE CAD layout

4.5 RF Amplifier

An Avago MGA-43228 RF power amplifier was chosen to increase the WPT demonstration system transmit power. At 2.45 GHz the MGA-43228 has approximately 30 dB of gain, an OP1dB of 35.5 dBm, and a 16.1% PAE. A functional block diagram is shown in Fig. 4.12. The MGA-43228 is a three-stage power amplifier based on Avago's proprietary 0.25 μm GaAs E-pHEMT (Enhancement-mode pseudomorphic high electron mobility transistor) process [32]. The MGA-43228 also includes shutdown, switchable gain, and power detection features. It is packaged in a 5 by 5 mm QFN chip with a ground paddle.

Fig. 4.12 Avago RF amplifier MGA-43228 pin configuration and internal block diagram [32]

The MGA-43228 has a specific turn-on and turn-off procedure to prevent damage to the amplifier. A higher voltage at the V_C pins than at the V_{BIAS} pin will cause a high current DC short at the V_C pins. The turn-on and turn-off sequence is shown in Fig. 4.13. The final step in the turn-on procedure where bias is applied to V_{BYP} is only used when the low-gain mode is desired.

Fig. 4.13 Avago RF amplifier MGA-43228 turn-on and turn-off sequence for DC power pins
[32]

The EAGLE CAD schematic and layout for the MGA-43228 is shown in Fig. 4.14 and Fig. 4.15, respectively. DC blocking capacitors C325 and C328 are located on the RF input and output nets to prevent DC voltage from the amp getting into other components in the RF path. Tuning capacitors C326, C327A, and C327B are included to optimize the return loss. Resistors R302 through R304 are responsible for setting the bias to the three amplifier stages. Bypass and filtering components are included on the supply lines and control lines to prevent unwanted RF feedback.

Fig. 4.14 Avago RF amplifier MGA-43228 EAGLE CAD schematic

Fig. 4.15 Avago RF amplifier MGA-43228 EAGLE CAD layout

4.6 Microwave Rectifier

A Schottky diode voltage doubler is used to convert incoming RF signals to a DC voltage. The voltage doubler is based on an Avago HSMS-2822 surface mount RF Schottky barrier diode. A voltage double is a combination of a clamper circuit and a detector [33]. The voltage of an incoming sine wave, which is symmetric around zero, is raised by the clamper so that the minimum voltage is zero. The input to the detector is the sine wave plus a DC offset equal to the peak voltage. The detected voltage is therefore the peak-to-peak voltage of the sine wave, double the peak amplitude detected by a single diode.

The HSMS-2822 Schottky diode, matching network, and voltage doubler were modeled using Agilent ADS, shown in Fig. 4.16 through Fig. 4.18, respectively. The package parasitic inductances and capacitances were found using Avago's application note: *Linear Models for Diode Surface Mount Packages* [34]. A lumped component matching network was designed to transform the impedance of the HSMS-2822 model to the 50 Ω characteristic impedance of the system. A three-component matching network was chosen because physically realizable components weren't possible using a two-component approach. Modelithics models were used for the components inside the matching network to ensure the model would accurately predict the behavior of the constructed circuit. All transmission line lengths and width transitions from the layout were accounted for in the microwave rectifier model.

Fig. 4.16 Avago Schottky Diode HSMS-2822 Agilent ADS Model

Fig. 4.17 Microwave Rectifier Matching Network

Microwave Rectifier Impedance

m1
freq=2.450GHz
S(1,1)=0.041 / 54.736
impedance = $Z_0 * (1.047 + j0.071)$

Fig. 4.19 Microwave Rectifier Input Impedance Agilent ADS Simulation

Microwave Rectifier Reflection Coefficient Magnitude

m1
freq=2.446GHz
dB(S(1,1))=-34.670
Min

m2
freq=2.450GHz
dB(S(1,1))=-27.646

Fig. 4.20 Microwave Rectifier Agilent ADS Simulation of Reflection Coefficient Magnitude

Microwave Rectifier Output Voltage vs. Input Power

Fig. 4.21 Microwave Rectifier Agilent ADS Simulation of Output Voltage vs. Input Power

Fig. 4.22 Microwave rectifier EAGLE CAD schematic

Fig. 4.23 Microwave Rectifier EAGLE CAD layout

4.7 Wireless Power Demonstration System

The final layout of all the circuits in the WPT demonstration system is shown in Fig. 4.24 through Fig. 4.28. The board is 4.75 inches long and 1.35 inches wide. Edge mount SMA connectors and semi-rigid coax cable were chosen to connect the circuits together so each component could be characterized individually.

Fig. 4.24 WPT demonstration system EAGLE CAD layout with visible top layer, vias, silk screen, and solder stop mask

Fig. 4.25 WPT demonstration system EAGLE CAD layout with visible layer 2 (GND), vias, silk screen, and solder stop mask

Fig. 4.26 WPT demonstration system EAGLE CAD layout with visible layer 3 (+5V, Signals), vias, silk screen, and solder stop mask

Fig. 4.27 WPT demonstration system EAGLE CAD layout with visible bottom layer (GND), vias, silk screen, and solder stop mask

Fig. 4.28 WPT demonstration system EAGLE CAD layout with all layers visible

5 Test Results

The printed circuits boards for the MPA, S-SRR strips and WPT demonstration system were constructed at Advanced Circuits. Two MPAs (SN1 and SN2), ten S-SRR strips, and two WPT demonstration system boards (SN1 and SN2) were ordered. An extra S-SRR strip and an extra WPT demonstration system board were ordered for redundancy as a precautionary measure. The finished WPT demonstration system, S-SRR strip, and MPA are shown in Fig. 5.1, Fig. 5.2, and Fig. 5.3, respectively. A parts list can be found in Table D.2

Fig. 5.1 WPT demonstration system

Fig. 5.2 S-SRR array PCB strip

Fig. 5.3 Microstrip Patch Antenna

5.1 MPA and Metamaterial Enhanced MPA

The MPA and MTM enhanced MPA were tested in Anechoic Chamber 2B at Raytheon Electronic Warfare Systems, shown in Fig. 5.4a and Fig. 5.4b, respectively. A foam structure was constructed to hold the S-SRR array PCB strips. Nylon rods with #10-32 threading were used to hold the MTM lens at specified distances away from the MPA.

Fig. 5.4 (a) MPA test setup (b) MTM enhanced MPA test setup

The measured data for MPA SN1 and SN2 is compared against the simulated characteristics in Fig. 5.5 through Fig. 5.9 and summarized in Table 5.1. The resonant frequency for both antennas is less than 1% different than the simulated value. Measured peak boresight gain has less than 0.2 dB error in comparison to the simulated value. The input impedance of SN1 and SN2 differs greatly from the simulated value due to the shift in resonant frequency. The half-power beam width in the E-plane and H-plane is larger than the simulated value by approximately 11 degrees for both units.

Fig. 5.5 Measured and simulated MPA boresight swept gain

Fig. 5.6 MPA measured and simulated input reflection coefficient

Fig. 5.7 MPA measured and simulated input impedance (real)

Fig. 5.8 MPA measured and simulated input impedance (imaginary)

Fig. 5.9 MPA SN1 measured and simulated radiation patterns at 2.45 GHz

Table 5.1 Simulated and measured MPA performance

	HFSS Optimized MPA Design	SN1 Measured	SN1 % Difference from Simulation	SN2 Measured	SN2 % Difference from Simulation
<i>f_r</i> (GHz)	2.454	2.474	0.81	2.468	0.57
Min S11 (dB)	-43.3	-37.3	13.86	-43.5	0.46
Peak Boresight Gain (dB)	2.9	2.75	5.17	2.71	6.55
MPA Port Z at 2.45 GHz [real] (Ω)	45.2	44.3	1.99	52	15.04
MPA Port Z at 2.45 GHz [imag] (Ω)	2.29	29.8	1201.31	23.6	930.57
E-Plane HPBW (deg)	99	110	11.11	109	10.10
H-Plane HPBW (deg)	74	83	12.16	83	12.16

The GRIN MTM lens was attached to the MPAs and the location was swept to optimize the boresight swept gain. The peak boresight swept gain was found when the GRIN MTM lens was located at 2.75 inches away from each MPA, which is the same location found in the HFSS simulation. The measured MTM swept gain optimization is shown in Fig. 5.10 and Fig. 5.11. The comparison of measured and simulated peak boresight swept gain is shown in Fig. 5.12. The peak boresight swept gain for MTM enhanced SN1 and MTM enhanced SN2 is 9.62 dBi and 9.77 dBi, respectively. Since the peak boresight swept gain of MTM enhanced SN2 is larger than MTM enhanced SN1, the GRIN MTM lens was only applied to SN2 for the remainder of the testing. The measured and simulated characteristics of the MTM enhanced MPA are shown in Fig. 5.13 through Fig. 5.17 and summarized in Table 5.2 and Table 5.3.

Fig. 5.10 MTM enhanced MPA SN1 measured boresight swept gain optimization

Fig. 5.11 MTM enhanced MPA SN2 measured boresight swept gain optimization

Fig. 5.12 MTM enhanced MPA measured and simulated boresight swept gain at 2.75 inches

Fig. 5.13 MTM enhanced MPA measured and simulated input reflection coefficient

Fig. 5.14 MTM enhanced MPA measured and simulated input impedance (real)

Fig. 5.15 MTM enhanced MPA measured and simulated input impedance (imaginary)

Fig. 5.16 MTM enhanced MPA SN2 measured and simulated radiation patterns 2.45 GHz

Table 5.2 MTM enhanced MPA measured and simulated performance comparison

	HFSS Optimized MTM MPA Design	MTM MPA SN2 Measured	MTM MPA SN2 % Difference
f_r (GHz)	2.445	2.457	0.49
Min S11 (dB)	-18.8	-16.76	10.85
Peak Boresight Gain (dB)	8.81	9.77	10.90
MPA Port Z at 2.45 GHz [real] (Ω)	42.4	40.19	5.21
MPA Port Z at 2.45 GHz [imag] (Ω)	-8.23	12.96	257.47
E-Plane HPBW (deg)	42	40	4.76
H-Plane HPBW (deg)	44	40	9.09

The measured resonant frequency for MTM enhanced MPA SN2 is less than 1% different in comparison to the simulated value. The measured peak boresight gain is greater than the simulated value by 0.96 dB. The measured input impedance differs greatly from the simulated value due to the shift in resonant frequency. The half-power beam width in the E-plane and H-plane has less than 10% error.

Fig. 5.17 MTM enhanced MPA SN2 and MPA SN2 measured radiation patterns

Table 5.3 Measured MPA and MTM enhanced MPA performance comparison

	MPA SN2 Measured	MTM MPA SN2 Measured	MTM MPA SN2 %Difference from MPA
<i>f_r</i> (GHz)	2.468	2.457	0.45
Min S11 (dB)	-43.5	-16.76	61.47
Peak Boresight Gain (dB)	2.71	9.77	260.52
MPA Port Z at 2.45 GHz [real] (Ω)	52	40.19	22.71
MPA Port Z at 2.45 GHz [imag] (Ω)	23.6	12.96	45.08
E-Plane HPBW (deg)	109	40	63.30
H-Plane HPBW (deg)	83	40	51.81

As shown in Table 5.3, adding the GRIN MTM lens increases the MPA gain by 7.06 dB. The MTM lens dramatically increases the directivity of the antenna, with the half-power beam width more than halved for the E-plane and H-plane. The resonant frequency shifts slightly lower with the MTM lens, by 9 MHz. The input impedance becomes slightly mismatched to the 50 Ω system when the lens is applied, degrading the reflection coefficient.

5.2 Voltage Controlled Oscillator

The Mini-Circuits ROS-2625-119+ VCO measured performance data is shown in Fig. 5.18 through Fig. 5.21. The measured tuning slope of 100 MHz/V matches well with the advertised tuning slope of approximately 101 MHz/V. The measured output power corresponds to the advertised typical output power of +4 dBm, but decreases slightly to approximately 3.5 dBm for tuning voltages above 2.65 V. The measured harmonic power levels are within the advertised specification of -20 dBc.

Fig. 5.18 Mini-Circuits ROS-2625-119+ tuning voltage vs. output frequency

Fig. 5.19 Mini-Circuits ROS-2625-119+ VCO output frequency vs. output power

Fig. 5.20 SN1 VCO tuned to 1.85V (~2.45 GHz), full-spectrum

Fig. 5.21 SN2 VCO tuned to 1.85V (~2.45 GHz), full-spectrum

5.3 Voltage Controlled Attenuator

The RFMD RFSA2013 voltage controlled attenuator performed poorly in all bench tests. The measured performance data is shown in Fig. 5.22 through Fig. 5.27. The measured minimum insertion loss of 4.65 dB at 2.45 GHz exceeds the advertised maximum specification of 3.5 dB minimum insertion loss by 1.15 dB. The advertised typical return loss of 15 dB differs from the worst-case measured input and output return loss of 6.5 dB and 7.1 dB at 2.45 GHz by 8.5 dB and 7.9 dB, respectively. The VCA was not used in the WPT transmit test due to the meager measured performance of the device.

Fig. 5.22 RFMD variable attenuator RFSA2013 transmission coefficient vs. control voltage for SN1

RFMD RFSA2013 SN2 Transmission Coefficient vs. Control Voltage

Fig. 5.23 RFMD variable attenuator RFSA2013 transmission coefficient vs. control voltage for SN2

RFMD RFSA2013 SN1 Input Reflection Coefficient vs. Control Voltage

Fig. 5.24 RFMD RFSA2013 VCA input reflection coefficient vs. control voltage for SN1

RFMD RFSA2013 SN2 Input Reflection Coefficient vs. Control Voltage

Fig. 5.25 RFMD RFSA2013 VCA input reflection coefficient vs. control voltage for SN2

RFMD RFSA2013 SN1 Output Reflection Coefficient vs. Control Voltage

Fig. 5.26 RFMD RFSA2013 VCA output reflection coefficient vs. control voltage for SN1

Fig. 5.27 RFMD RFSA2013 VCA output reflection coefficient vs. control voltage for SN2

5.4 Ceramic Band Pass Filter

The test results for the Johanson Technology 2450BP39D100C ceramic band pass filter did not meet the specifications advertised in the data sheet. The measured performance data is shown in Fig. 5.28 through Fig. 5.30. The measured minimum insertion loss of 1.5 dB was 0.3 dB greater than the advertised 1.2 dB maximum value for the minimum insertion loss. The measured 3 dB bandwidth of 480 MHz differs from the advertised bandwidth of 100 MHz by 380 MHz. The measured 3 dB pass band occurred between 2.35 GHz to 2.83 GHz, in comparison to the advertised pass band from 2.44 GHz to 2.45 GHz. The minimum measured input and output return loss was approximately 6 dB in the desired frequency band, 3.5 dB lower than the advertised return loss of 9.5 dB. The JTI BPF was used in the WPT transmit test, regardless of the poor performance, to aid in harmonic rejection.

Fig. 5.28 Johanson 2450BP39D100C BPF transmission coefficient

Fig. 5.29 Johanson 2450BP39D100C BPF input reflection coefficient

Fig. 5.30 Johanson 2450BP39D100C BPF output reflection coefficient

5.5 RF Amplifier

The specifications for minimum gain and minimum return loss advertised by Avago for the MGA-43228 RF power amplifier are contradicted in the data sheet. Avago claims in the electrical specifications portion of the data sheet to have minimum gain and minimum return loss of 35 dB and 10 dB, respectively. The S-parameter portion of the data sheet lists them as 30 dB and 1.18 dB, respectively. Considering these contradictions, the MGA-43228 units under test were evaluated on desired performance and not whether they met the manufactures' specifications. Measured performance data is shown in Fig. 5.31 through Fig. 5.39. The measured gain at 2.45 GHz for SN1 and SN2 was 31.7 dB and 32.9 dB, respectively. The measured input return loss at 2.45 GHz for SN1 and SN2 was 5.79 dB and 5.50 dB, respectively. The measured output return loss

at 2.45 GHz for SN1 and SN2 was 3.38 dB and 2.92, respectively. The measured 1 dB compression point at 2.45 GHz for SN1 and SN2 was 34.8 dBm and 34.7 dBm, respectively. SN2 was the ideal candidate to be the transmitter for the WPT transmit test in comparison to SN1 because of its higher gain.

Fig. 5.31 Avago MGA-43228 measured and published gain

Fig. 5.32 Avago MGA-43228 measured and published input reflection coefficient

Fig. 5.33 Avago MGA-43228 measured and published output reflection coefficient

Fig. 5.34 Avago MGA-43228 SN1 input power vs. gain

Fig. 5.35 Avago MGA-43228 SN2 input power vs. gain

Fig. 5.36 Avago MGA-43228 input power vs. gain at 2.45 GHz

Fig. 5.37 Avago MGA-43228 SN1 input power vs. output power

Fig. 5.38 Avago MGA-43228 SN2 input power vs. output power

Fig. 5.39 Avago MGA-43228 input power vs. output power at 2.45 GHz

5.6 Microwave Rectifier

The measured microwave rectifier performance didn't match the simulations as anticipated. As shown in Fig. 5.40, rectifiers SN1 and SN2 are best matched to 50Ω at 2.56 GHz and 2.58 GHz, respectively. The measured and simulated input impedance have drastically different characteristics, shown in Fig. 5.41. The drastic impedance difference can be attributed to component value tolerances in the matching network and variation in package parasitics used in the Agilent ADS model.

Fig. 5.40 Microwave Rectifier measured and simulated input reflection coefficient

Microwave Rectifier Impedance

Fig. 5.41 Microwave rectifier measured and simulated input impedance

Despite the poor impedance match of the rectifiers, the measured output voltage is nearly identical to the simulation. The measured output voltage versus input power is shown in Fig. 5.42 through Fig. 5.44. Surprisingly, the peak output voltage occurred at a frequency that was higher than the peak return loss. As shown in Fig. 5.44, SN1 outperforms SN2 at 2.45 GHz and therefore it is the clear choice to be the receiver in the WPT transmit test.

Fig. 5.42 Microwave rectifier SN1 measured input power vs. output voltage

Fig. 5.43 Microwave rectifier SN2 measured input power vs. output voltage

Fig. 5.44 Microwave rectifier measured comparison at 2.45 GHz

Fig. 5.45 Microwave Rectifier SN2 Output Voltage Ripple with 14 dBm input at 2.45 GHz

5.7 System Demonstration

A full system demonstration was performed in Anechoic Chamber 2B at Raytheon Electronic Warfare Systems in Goleta. A block diagram of the test setup is shown in Fig. 5.46. A two-foot transmit distance was chosen to ensure that enough power reached the microwave rectifier with and without the GRIN MTM lens. As mentioned previously, the voltage controlled attenuator was omitted from the final test due to poor performance.

Fig. 5.46 WPT transmit test block diagram

Images of the test setup are shown in Fig. 5.47 through Fig. 5.49. RF absorber was used to minimize reflections and the exposed metal from the test cart was verified to not impact test results. The test was conducted by transmitting a signal using WPT demonstration system SN2 and receiving it with WPT demonstration system SN1. The test was repeated using a spectrum analyzer to verify the power level and frequency of the transmitted signal. Fig. 5.48 shows the receive antenna and spectrum analyzer on the test cart. Fig. 5.49 shows the transmitter and MTM enhanced MPA.

Fig. 5.47 WPT transmit test setup

Fig. 5.48 WPT transmit test: receiver with spectrum analyzer

Fig. 5.49 WPT transmit test: transmitter with MTM enhanced MPA

Screen captures of the received signal seen by the spectrum analyzer with and without the GRIN MTM lens are shown in Fig. 5.50 and Fig. 5.51, respectively. An average power gain of 8.15 dB is demonstrated between 2.42 GHz and 2.51 GHz by adding the GRIN MTM lens. This power gain is 1.15 dB higher than the expected power gain of the GRIN MTM lens and could be due to misalignment of the antennas when the MTM was removed. To ensure that harmonics aren't contributing to the received power, Fig. 5.52 shows the harmonic purity of the transmitted signal.

Fig. 5.50 WPT demonstration system test with MTM enhanced MPA

Fig. 5.51 WPT demonstration system test without MTM

Fig. 5.52 WPT system test with MTM enhanced MPA, full spectrum

The expected received power was calculated using Friis' formula shown in equation (5.1) [35]. To solve for the received power P_r , the transmitted power P_t , receive antenna gain G_r , transmit antenna gain G_t , transmit distance R , and wavelength λ must be determined. The transmit and receive antenna gain were approximated using a 6th-order polynomial trend fit of the measured antenna swept gain so that the gain can be calculated for any transmit frequency, seen in Fig. 5.53. The transmit power and frequency were determined using the spectrum analyzer and sweeping the VCO frequency across the desired range. The transmit distance is 2 feet, as shown in the test setup block diagram.

$$P_r(dB) = G_r(dB) + G_t(dB) + P_t(dBm) - 20 \log\left(\frac{4\pi R}{\lambda}\right) \quad (5.1)$$

Fig. 5.53 Swept gain 6th order polynomial trend fit

The measured and expected received power levels are shown in Fig. 5.54.

The average error between the measured and expected received power with and without the GRIN MTM lens is 0.638 dBm and 1.62 dBm, respectively. The effective isotropic radiated power was calculated using equation (5.2) and is shown in Fig. 5.55.

$$EIRP = P_t(dBm) + G_t(dB) \quad (5.2)$$

Fig. 5.54 Measured and expected received power

Fig. 5.55 Calculated EIRP with and without MTM lens

Finally, the measured microwave rectifier output voltage with and without the GRIN MTM lens is shown in Fig. 5.56. The average increase in output voltage across the frequency band is 1.34 V. It is therefore shown that the GRIN MTM lens significantly increases the received power in the WPT demonstration system.

Fig. 5.56 Measure microwave rectifier output voltage with and without MTM lens

6 Conclusion

This project successfully demonstrated the performance enhancement of a gradient refractive index metamaterial lens when applied to a wireless power transmission system. A WPT demonstration system, MPA, and GRIN MTM lens were designed, built, and tested herein. The theoretical and measured gain improvement of the MPA due to the GRIN MTM lens is 5.91 dB and 7.06 dB, respectively. The measured resonant frequency for the MTM enhanced MPA SN2 is less than 1% different in comparison with the simulated value. The measured peak boresight gain is greater than the simulated value by 0.96 dB. The measured half-power beam width in the E-plane and H-plane has less than 10% error with respect to the simulated values.

Ansoft's High Frequency Structure Simulator was used to model the performance and behavior of a microstrip patch antenna, S-shaped split ring resonator, and GRIN MTM lens. The powerful optimization capability of HFSS was instrumental in maximizing the performance of this design. The field and vector overlays proved crucial in understanding the focusing mechanism behind the GRIN MTM lens.

Inexpensive commercial-off-the-shelf components were characterized and effectively integrated to create a WPT demonstration system. Even though most of the components didn't behave in an ideal manner, the system was able to output 32 dBm at 2.45 GHz. An average power gain of 8.15 dB was demonstrated between 2.42 GHz and 2.51 GHz by adding the GRIN MTM lens. This power

gain is 1.15 dB higher than the expected power gain of the GRIN MTM lens and was due to misalignment of the antennas when the MTM was removed. The average error between the measured and expected received power with and without the GRIN MTM lens is 0.638 dBm and 1.62 dBm, respectively.

Despite the success of this project, there are many research opportunities available to improve and characterize MTM lenses. This design ultimately resulted in a GRIN lens, but a negative refractive index (NRI) lens or a lens with refractive index near zero could have been used. The power handling capability of MTM lenses is also a relatively untouched subject with potential applications for radar or satellite communications. Further investigation of metamaterials is necessary to fully realize the benefits of these unique structures.

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APPENDIX A : MPA Design

```
% Rectangular Microstrip Patch Antenna
% Travis Heffernan
% Cal Poly Master's Thesis

% This program calculates the dimensions of a rectangular patch
antenna
% using user-specified dielectric relative permeability
epsilon_r,
% dielectric height h, and desired resonant frequency fr.

clear
clc
close all

% Constants
mu_0 = 1.25663706e-6; %Permeability of free space
epsilon_0 = 8.85418782e-12; %Permittivity of free space
epsilon_r = 4.4; %Relative Permittivity of FR-4
C = 2.99792458e8; %Speed of light
fr = 2.45e9; %Resonant frequency

% Patch Dimensions
h = 59*2.54e-5; %Height of dielectric, meters
W = (C/(2*fr))*sqrt(2/(epsilon_r+1)); %Patch width, meters
epsilon_eff = (epsilon_r+1)/2 +...
    ((epsilon_r-1)/2)*(1+12*h/W)^(-1/2); %Effective dielectric
constant
del_L = 0.412*h*((epsilon_eff+0.3)*(W/h+0.264))/...
    ((epsilon_eff-0.258)*(W/h+0.8)); %Length caused by fringing
fields
L = C/(2*fr*sqrt(epsilon_eff))-2*del_L; %Patch length, meters
Leff = L+2*del_L; %Effective patch length, meters
Leff_mil = Leff/2.54e-5; %Effective patch length, mils
del_L_mil = del_L/2.54e-5; %Length caused by fringing fields,
mils
W_mil = W/2.54e-5 %Patch width, mils
L_mil = L/2.54e-5 %Patch length, mils

% Patch Input Impedance
Zin = 90*epsilon_eff^2/(epsilon_eff-1)*(Leff/W)^2;

% 1/4 Wave Transformer Impedance
Zin_qw = sqrt(Zin*50)
```

APPENDIX B : S-SRR Design

```
% S-SRR Characterization
% Travis Heffernan
% Cal Poly Master's Thesis

clear
clc
close all

% Constants
mu_0 = 1.25663706e-6; %Permeability of Free Space
epsilon_0 = 8.85418782e-12; %Permittivity of Free Space
epsilon_r = 4.4; %Relative Permittivity of PCB Dielectric

% Frequency Sweep
freq = (0:0.1:6); %Frequency in GHz
omega = freq*1e9*2*pi; %Frequency in rad/s

% S-SRR Parameters
a = 1400*2.54e-5;%Unit Cell Dimension in the X-direction, mils-
>meters
b = 0.5*a; %Unit Cell Dimension in the Z-direction
h = 0.4*a; %S-SRR Height
w = 0.75*a; %S-SRR Width
c = a/20; %Trace width
d = 59*2.54e-5; %Distance between S-Shaped structures, mils to
meters
l = 500*2.54e-5;%Unit cell Dimension in the Y-direction, mils to
meters
F = 0.3; %Fractional Volume of Each Loop
S = a*b; %Area of Unit Cell in the XZ Plane
sigma_s = 0.5; %Resistance of Metallic Strips

% Capacitance Between S-SRR Strips
Cs = epsilon_r*epsilon_0*((h*c)/d) + epsilon_0*((h*c)/(l-d));
Cm = Cs;

% Parameter Simplifications
A_sigma = omega*mu_0*S*(F^2+F^2)*sigma_s*l;
B_sigma = (sigma_s*l)^2;
C_sigma = (omega*mu_0*S*(F+F)-(2./omega)*(l/Cs+l/Cm))*sigma_s*l;
X = (omega*mu_0*F*S).^2.*(1-l./(omega*mu_0*F*S*Cs));
D_sigma = A_sigma./X;
E_sigma = B_sigma./X;
G_sigma = C_sigma./X;
m = F/F;
n = Cm/Cs;
```

```

% Calculation of Magnetic Permeability
mu_eff = 1-(2*F+1i*D_sigma)./(1-(1./(omega.^2*mu_0*F*S))*...
    (1/Cs+(2*1)/Cm)-E_sigma+1i*G_sigma);

% Calculation of Magnetic Resonant Frequency and Magnetic Plasma
Frequency
omega_m0 = sqrt((1/(mu_0*F*S))*(1/Cs+(2*1)/Cm));
f_m0 = omega_m0/(2*pi);
omega_mp = omega_m0*sqrt(1/(1-2*F));
f_mp = omega_mp/(2*pi);

% Plot Permeability
figure()
plot(freq,real(mu_eff),freq,imag(mu_eff));
axis([0 6 -5 5])
xlabel('Frequency (GHz)')
ylabel('Effective Permeability, \mu_e_f_f')
title('S-SRR Effective Permeability Frequency
Response','fontsize',16)
grid on
legend('Real','Imaginary')

```

APPENDIX C : Parameter Extraction

```
% S-SRR Metamaterial Parameter Extraction
% Travis Heffernan
% Cal Poly Master's Thesis

clear
clc
close all

% Import Data from HFSS
temp1=csvread('array890_S11.csv',1,0); %Frequency, Real S11,
Imaginary S11
freq=temp1(:,1); %Frequency in GHz
S11 = temp1(:,2)+1j*temp1(:,3);
temp2=csvread('array890_S21.csv',1,0); %Frequency, Real S21,
Imaginary S21
S21 = temp2(:,2)+1j*temp2(:,3);

%Constants
C = 299792458; %Speed of light
d = 500*2.54e-5; %S-SRR unit cell slab thickness, mils to meters
k0 = 2*pi*freq*1e9/C; %Free space wave number

% Parameter Calculations
V1 = S21+S11;
V2 = S21-S11;
mu = 2*(1-V2)./(1j*k0*d).*(1+V2));
epsilon = mu + 1j*2*S11./(k0*d);
% epsilon = 2*(1-V1)./(1j*k0*d).*(1+V1));
n = sqrt(mu.*epsilon);
wave_z_sqr = ((S11+1).^2-S21.^2)./((S11-1).^2-S21.^2);
wave_z = sqrt(wave_z_sqr);
gamma = (wave_z-1)./(wave_z+1);

for i = 1:length(n)
    if real(epsilon(i))<0 & real(mu(i))<0
        n(i) = -n(i);
    end
end

figure()
plot(freq,real(mu),freq,imag(mu),freq,real(epsilon),freq,imag(eps
ilon),...
    freq,abs(mu),freq,abs(epsilon));
title('Extracted Permeability and Permittivity of S-SRR')
xlabel('Frequency (GHz)')
ylabel('unitless')
legend
('Real(\mu)', 'Imag(\mu)', 'Real(\epsilon)', 'Imag(\epsilon)',...
    'Magnitude(\mu)', 'Magnitude(\epsilon)' )
grid on
```

```

figure()
plot(freq,real(n),freq,imag(n),freq,abs(n));
title('Extracted Refractive Index')
xlabel('Frequency (GHz)')
ylabel('unitless')
legend ('Real(n) ','Imag(n) ','Magnitude(n) ')
grid on

figure()
plot(freq,real(wave_z),freq,imag(wave_z),freq,abs(wave_z));
title('Extracted Wave Impedance')
xlabel('Frequency (GHz)')
ylabel('\Omega')
legend ('Real(\eta) ','Imag(\eta) ','Magnitude(\eta) ')
grid on

figure()
plot(freq,real(gamma),freq,imag(gamma),freq,abs(gamma));
title('Reflection Coefficient')
xlabel('Frequency (GHz)')
ylabel('unitless')
legend ('Real(\Gamma) ','Imag(\Gamma) ','Magnitude(\Gamma) ')
grid on

```

APPENDIX D : Analysis of Senior Project Design

D.1 Summary of Functional Requirements

This project demonstrates that metamaterials (MTMs) can improve antenna gain and thereby increase microwave wireless power transmission (WPT) system performance. A MTM lens and simple WPT demonstration system were designed herein to operate at 2.45 GHz with an effective isotropic radiated power (EIRP) above 36 dBm. The functional requirements are listed in Table D.1.

Table D.1 WPT system functional requirements

Operating Frequency	2.45 GHz
EIRP	>36 dBm
MTM Gain Improvement	>0 dB

D.2 Primary Constraints

The main constraint with this project was finding computer hardware that could handle the MTM array HFSS simulation. As discussed in section 3.3, a quad processor server with 256 GB of RAM had difficulty with the parametric analysis used to optimize the MTM array.

Another limitation this project overcame was achieving desired performance with low cost PCB material. The MPA, MTM enhanced MPA, and WPT demonstration system were originally designed using Advanced Circuit Materials from Rogers Corporation [36]. RO4003C^{TM} was originally used for the WPT demonstration system and RT/duroid® 5880 was originally used for the MPA and MTM enhanced MPA. The production estimate from Advanced Circuits, where the PCBs were manufactured, was approximately 3-times the cost of using FR-4.

D.3 Economic Impact

WPT can provide energy to areas where wired networks are impractical or impossible due to physical geography or lack of infrastructure. WPT therefore provides a new market for electric utility companies where wired networks weren't economically feasible. Solar power satellites have the ability to point the energy at a desired location for an allotted amount of time, which could allow electric utility companies to provide wireless power to numerous remote locations on a timeshare basis. This would be greatly beneficial for natural disaster relief, impoverished nations, and remote or temporary human settlements (military or otherwise).

The total cost associated with this project is \$745.46, shown in . The initial estimated cost proposed to Raytheon in November of 2011 was \$10,000. See Fig. D.1 and Fig. D.2 for project timelines. The equipment costs are excluded from this project, but are estimated at over \$1 million USD for the vector network analyzer, spectrum analyzer, and anechoic chamber. The materials are the main cost of this project and they accrue towards the middle of the production cycle, after the design phase but before the testing phase. The project benefits are realized after the testing phase, once the product is complete. Project earnings are related to the power handling capability of the system minus the cost to produce. The average "On Peak" price for electricity for California, Nevada, and Oregon in December 2012 is \$37.16 per megawatt hour [37]. A company that produces a large-scale system with a 500 MW power handling capability would therefore earn \$18,580 per hour.

The general cost of improving a WPT system with a MTM lens is related to the aperture size of the transmit antenna and the chosen MTM structure. A large transmit antenna aperture will increase the directivity of the radiated power and therefore a smaller MTM lens can be used, and vice versa. The costs associated with a large antenna or large MTM lens must be considered and balanced when manufacturing a large-scale system. A commercial or military system based on this design could cost over 1000-times more than the cost associated with this project. This cost estimate is based on expected labor costs, material costs, test equipment costs, and facility overhead. Labor cost estimates include PCB manufacturing labor, test engineer labor, and project management labor. Material costs would increase significantly for a commercial or military system since higher power handling capabilities are needed to make WPT a viable power distribution option. Facility overhead is also a significant factor for increasing the

cost of a large-scale system since manufacturing facilities are expensive to occupy and maintain.

Table D.2 Final project cost

QTY	Ref	Part Description	Manufacturer	Part Number	Supplier	Cost Per	QTY Cost	Package
16	C101, C103, C201, C202, C209, C210, C402, C403	CAP, 1000pF, 10%, 25V, X7R	Murata	GRM155R71H102KA01D	Digikey	\$0.02	\$0.30	0402
	C102, C104, C204, C211	CAP, 1uF, 10%, 25V, X7R	TDK Corporation	C2012X7R1E105K	Digikey	\$0.16	\$1.28	0805
	C301, C305, C311, C322	CAP, 0.1uF, 10%, 16V, X7R	Murata	GRM155R71C104KA88D	Digikey	\$0.02	\$0.17	0402
2	C304	CAP, 8.2pF, ± 0.5 pF, 50V, COG	Murata	GRM1885C1H8R2DZ01D	Digikey	\$0.08	\$0.16	0603
8	C307, C313, C325, C328	CAP, 7.5pF, ± 0.5 pF, 50V, COG	Murata	GRM1555C1H7R5DZ01D	Digikey	\$0.04	\$0.35	0402
	C308	CAP, 2.4pF, ± 0.25 pF, 50V, COG	Murata	GJM1555C1H2R4CB01D	Digikey	\$0.11	\$0.21	0402
	C309, C329	CAP, 22uF, 10%, 16V, X5R	TDK Corporation	C2012X5R1C226K	Digikey	\$0.77	\$3.08	0805
2	C312	CAP, 2.2pF, ± 0.25 pF, 50V, COG	Murata	GJM1555C1H2R2CB01	Digikey	\$0.08	\$0.16	0402
2	C323	CAP, 0.022uF, 10%, 25V, X7R	Murata	GRM155R71E223KA61	Digikey	\$0.04	\$0.08	0402
2	C326	CAP, 0.4pF, ± 0.1 pF, 50V, COG	Kemet	CBR04C408B5GAC	Digikey	\$0.16	\$0.32	0402
2	C327a	CAP, 1.8pF, ± 0.25 pF, 50V, COG	Murata	GJM1555C1H1R8CB01D	Digikey	\$0.11	\$0.21	0402
2	C327b	CAP, 2.0pF, ± 0.25 pF, 50V, COG	Murata	GJM1555C1H2R0CB01	Digikey	\$0.11	\$0.21	0402
2	C401	CAP, 1.2pF, ± 0.25 pF, 50V, NP0	Murata	GJM1555C1H1R2CB01D	Digikey	\$0.10	\$0.20	0402
2	R205	RES, 100 Ω , 5%, 1/10W	Yageo	RC0603JR-07100RL	Digikey	\$0.01	\$0.02	0603
4	R302, R304	RES, 27k Ω , 1%, 1/10W	Yageo	RC0603FR-0727KL	Digikey	\$0.01	\$0.06	0603
2	R304	RES, 22k Ω , 5%, 1/10W	Yageo	RC0603JR-0722KL	Digikey	\$0.01	\$0.02	0603
2	R401	RES, 100k, 1%, 1/3W	ESR18	ESR18EZPF1003	Digikey	\$0.22	\$0.44	1206
2	L301	IND, 1nH, 5%	Coilcraft	0402HP-1N0XJLU	Coilcraft	\$1.12	\$2.24	0402
2	L401	IND, 6.2nH, 2%	Coilcraft	0402HP-2N4XGLU	Coilcraft	\$1.12	\$2.24	0402
2	L402	IND, 2nH, 2%	Coilcraft	0402HP-2N4XGLU	Coilcraft	\$1.12	\$2.24	0402
2	D401	DIODE, SCHOTTKY, 15V	Avago	HSM5-2822-TR1G	Digikey	\$0.99	\$1.98	SOT-23
2	U201	Variable Attenuator	RFMD	RFSA2013	RFMD	\$6.22	\$12.44	16-QFN
2	U301	RF Amp	Avago	MGA-43228	Avago	\$15.40	\$30.80	28-QFN
2	U501	BPF, 2.45GHz, 1.2dB IL	Johanson Technology	2450BP39D100CE	Digikey	\$0.59	\$1.18	1008
2	XO101	VCO, 2370-2600 MHz	Mini Circuits	ROS-2625-119+	Mini Circuits	\$19.95	\$39.90	CK605
8	J301	HEADER, 2 row, 4 pos., 100 mil	3M	961204-6404-AR	Digikey	\$0.24	\$1.92	
8	J311	JUMPER, 2 pos. 100 mil	Sullins Connector Solutions	QPC02SXGN-RC	Digikey	\$0.11	\$0.88	
18	J101, J201, J202, J301, J302, J401, J501, J502, J1	CONN JACK END LAUNCH PC GOLD SMA	Emerson	142-0701-851	Digikey	\$5.30	\$95.38	
	2	WPT Demonstration System	Advanced Circuits	demo_V5	Advanced Circuits	\$66.00	\$132.00	
	2	Microstrip Patch Antenna	Advanced Circuits	MPA_v2	Advanced Circuits	\$33.00	\$66.00	
10		S-SRR strips	Advanced Circuits	MTM_v2	Advanced Circuits	\$33.00	\$330.00	
8		#10-32 Nylon Machine Screw Nut (2-Pieces)	Home Depot	86868	Home Depot	\$0.62	\$4.96	
4		#10 Clear Nylon Washer (5-Pieces)	Home Depot	86938	Home Depot	\$0.61	\$2.44	
1		Nylon #10-32 Threaded Rod (.190 dia.)	U.S. Plastics Corp	91843	U.S. Plastics Corp	\$5.60	\$5.60	
1		Foam board for MTM structure	FloraCraft		Michael's	\$5.99	\$5.99	

Total Cost \$745.46

Fig. D.1 Estimated Project Timeline

Fig. D.2 Actual Project Timeline

D.4 If Manufactured on a Commercial Basis

The number of devices sold per year depends on the power handling capability of the systems and the industry demand. The devices constructed within this project were proof-of-concept and aren't a viable commercial product due to low efficiency COTS components. A large-scale efficient system with 500MW power handling capability would be more difficult and take longer to manufacture than this project, but the necessary manufacturing facilities could be scaled to accommodate any quantity. As mentioned in APPENDIX D.3, the cost of producing a large-scale system could cost over 1000-times the cost of this project. A case study and market research are necessary to determine the current demand for a WPT system and to give an accurate estimate for the expected number of devices per year. Typical power handling requirements are also needed to estimate the cost per device, purchase price, and expected profit. Current WPT market trends necessary to make such estimates are available for \$450 USD [38]. In 1999, however, costs for terrestrial WPT at distances up to tens of km were on the order of several million dollars per MW-km [39].

D.5 Environmental Impact

The natural resources and ecosystem services used by WPT systems is highly dependent upon the power source driving the system. WPT provides the conduit for energy transmission and therefore all benefits and drawbacks for any given power production method can be associated with it. WPT also uses metals and dielectric materials to fabricate the structures and PCBs necessary to transmit power. Electronic manufacturing chemicals and processes are caustic and harmful to the environment.

In addition to the environmental impacts contributed by the energy source and materials, WPT impacts the environment with large-aperture antennas necessary for collecting high power density microwave energy greater than 100 W/m^2 . The aperture size of the receive antenna can be on the order of a photovoltaic power station, which also carries all of the negative environmental impacts of a solar farm. Solar farms are widely known to displace animal populations, expose the environment to hazardous materials, and alter the landscape due to construction [40]. Microwave WPT can also expose biological matter to RF power, which can cause cellular deterioration due to excessive heat [41].

D.6   Manufacturability

The MPAs, MTM, and WPT demonstration system are all constructed using FR-4 PCB substrate and surface mount components. FR-4 is one of the most commonly used PCB substrates and is a manufacturing option available from every PCB manufacturer. Most PCB manufacturers now provide assembly services that include pick-and-place component placement and reflow soldering, which is ideal for surface mount devices. The only limitation of pick-and-place machines is that components must have vacant “keep out” regions surrounding them to ensure the robotic arm can maneuver the component into place. The MPA, MTM, and WPT demonstration system satisfy all “keep out” requirements for pick-and-place assembly.

D.7   Sustainability

The consideration of ecology, equity, and economy are at the crux of proper sustainable manufacturing practices. Even though WPT poses environmental risks, discussed in APPENDIX D.5, it can provide energy to remote regions using green energy sources such as solar, wind, hydroelectricity, and other alternative power production methods. Energy proliferation using timeshared solar power satellites, mentioned previously, could provide energy to third world countries at a discounted rate, which would dramatically increase the standard of living. Even at a discounted rate, timeshared solar power satellites would provide additional revenue for utility companies, therefore improving the economy.

Maintenance for MTM enhanced WPT system is minimal for the antennas and MTM array since they are passive metal structures. The transmitter and receiver contain all the active circuitry and could be difficult to maintain. High reliability components and a modular system design would significantly increase the expected lifespan of the system and minimize waste due to faulty components. A modular system design would also easily allow for upgrades to the power handling capability.

D.8   Ethical Issues

Engineering can radically impact the quality of life for all people. Engineers consequently have a responsibility to exemplify the pinnacle of morality and integrity. IEEE Policy Section 7.8 and NSPE Code of Ethics for Engineers discuss in detail the obligations and expected conduct of an engineer [42], [43]. The

fundamental canons of the NSPE Code of Ethics for Engineers dictate that engineers, in fulfillment of their professional duties shall [43]:

1. Hold paramount the safety, health, and welfare of the public.
2. Perform services only in areas of their competence.
3. Issue public statements only in an objective and truthful manner.
4. Act for each employer or client as faithful agents or trustees.
5. Avoid deceptive acts.
6. Conduct themselves honorably, responsibly, ethically, and lawfully so as to enhance the honor, reputation, and usefulness of the profession.

High power RF energy has a lethal effect on biological organisms and a misused or poorly designed WPT system could be an effective “death ray”. Ensuring proper alignment of large-scale MTM enhanced WPT systems would reduce the possibility catastrophic damage to biological organisms. The transmit side lobe and back lobe power levels should be minimized to eliminate radiation in an undesired direction and prevent RF exposure to biological organisms.

An ethical issue for WPT systems could arise if energy distribution were biased in favor of a particular region, ethnic group, or social class. Reserving energy from a WPT system for wealthy, Caucasian Americans would violate the code of ethics. Wireless power should be available for all people.

D.9 Health and Safety Concerns

Microwave power can be harmful to living organisms because it can heat biological matter. Excessive heat can cause cellular damage, teratogenic effects, sterility, and developmental effect when exposure limits exceed 4W/kg [41] . Animals, plants, and humans would be adversely affected if they were exposed to power densities necessary for large-scale WPT.

D.10 Social and Political Impact

WPT provides a new opportunity to give and trade energy between countries. Rather than providing aid in the form of gas generators to countries experiencing a natural disaster, it would be possible to beam the energy to the affected area. Countries without energy independence could also barter goods and services for wireless power.

A large-scale MTM enhanced WPT system could positively impact inhabitants of remote communities, military personnel, emergency personnel, politicians, and engineers. Inhabitants of remote communities could benefit from WPT if they previously didn't have access to electricity. Military and emergency personnel would benefit from WPT if they were in a location that didn't have access to the power grid. Politicians could benefit from WPT by setting up trade agreements to provide power to other countries. Engineers would benefit from WPT by the increased number of jobs necessary to create a large-scale system.

Globalization is the only negative socio-economic impact that a MTM enhanced WPT system would cause. Inhabitants of remote communities that previously didn't have electricity would be exposed to global culture and may lose their cultural identity. Sending engineers into a delicate cultural region may cause tension and incite rage from the inhabitants due to differing cultural and religious traditions.

D.11 Development

Ansoft's High Frequency Structure Simulator (HFSS) was used extensively throughout this project and was a tool that I've never used before. All analysis techniques and simulation tools within HFSS were learned using Ansoft's Online Help document and from Raytheon colleagues [11]. This project would not be possible without this powerful electromagnetic simulation software.

Agilent's Advanced Design System software was used for RF circuit simulation. This software was used to design the microwave rectifier and determine the microstrip trace widths. Even though I was already very familiar with this software, this project provided an avenue to hone my RF circuit design skills.

CadSoft's EAGLE PCB design software was used as the layout tool for the MPA, MTM, and WPT demonstration system circuit boards. Even though I was already very familiar with this software, this project provided an avenue to hone my RF PCB layout skills.

Constructing the circuit cards improved my hand-soldering ability since there were components using leadless QFN packages with ground paddles. Testing the MPA, MTM enhanced MPA, and WPT demonstration system improved my measurement skills using spectrum analyzers, vector network analyzers, and anechoic chambers.

D.12 Literature Search

The following is a reiteration of the thesis References section that includes applicable literature used within the Analysis of Sr. Project Design. This is included so the Analysis of Sr. Project Design appendix can be a standalone document.

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