HIGH-PERMITTIVITY HEMISPHERICAL LENS FOR MIMO APPLICATIONS WITH CLOSELY-SPACED ANTENNAS

by

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A thesis submitted in conformity with the requirements for the degree of Master of Applied Science
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Abstract

With the rapid adoption and development of new standards, Multiple-Input Multiple-Output (MIMO) technology is becoming a necessity in current wireless systems. One problem posed by using multiple antennas at a transmitter or receiver is the undesirable effect of signal correlation between closely-spaced radiating elements. This thesis presents the concept, design, and evaluation of a hemispherical lens antenna for use in MIMO systems. A high-permittivity dielectric material allows radiating elements to be placed in close proximity with reduced spatial correlation effects. An intermediate matching layer and a hemispherical lens design facilitate the preservation of the pattern characteristics in the transition between the dielectric and free-space. The antenna was simulated against benchmark antenna arrays in free-space and showed a 35%-70% improvement in channel capacity in multipath-rich environments, showing strength as a candidate for further development in MIMO applications.
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Chapter 1
Introduction

The area of wireless telecommunications deals with the transfer of information by electromagnetic means between multiple points without a physical connection. Developments in wireless telecommunications have been used in a variety of applications including radio, mobile telephones, wireless energy transfer, and wireless data communications. While current standards enable the possibility for wireless access to the internet and data transfer between devices (foregoing the traditional requirement for a wired connection), the supporting technologies are relatively nascent and frequently changing.

1.1 Development of Wireless Data Communications

Wireless local area network (WLAN) computer communication in the 2.4, 3.6, and 5 GHz frequency bands are governed by the IEEE 802.11 set of standards. The first release of the 802.11 legacy protocol in 1997 did not see widespread adoption due to the limited data rate of 1-2 Mbit/s, which was eclipsed by data rates of the new wired broadband alternatives of cable and digital subscriber line (DSL) technology. Two years later, the 802.11a/b protocols saw better deployment due to improvements in data rate and range and became the standard for WLANs. The 802.11g protocol, which operated in the 2.4 GHz band like 802.11b but applied the modulation scheme of 802.11a, further improved the viability of WLAN technology and was quickly adopted well before ratification in June 2003 due to the overwhelming desire among consumers for higher data rates. It was not until 2009 that the current 802.11n protocol was
released, promising further increases to WLAN range and data rates up to three times better than 802.11g [1].

Meanwhile, the proliferation of mobile phones and rapid improvement in available computing power on portable devices has led to the increased use of phones and tablet computers for mobile internet access. In addition to support for the 802.11 family of Wi-Fi protocols, these products have also used a variety of mobile technologies to access the internet. Early mobile internet access using the EDGE (GSM) and EV-DO (CDMA) standards provided low data rates similar in performance to the 802.11-1997 protocol, but were sufficient given the limited capabilities of the phones that used them. The introduction of the third generation (3G) Universal Mobile Telecommunications System (UMTS) was comparable to the 802.11a/b upgrade and allowed for more versatile internet-enabled applications on mobile devices. The introduction of the Evolved High-Speed Packet Access (HSPA+) and Long Term Evolution (LTE) standards enabled the possibility for mobile internet access at speeds comparable to broadband internet, and the current development of further improved fourth generation (4G) technologies like 802.16 (WiMAX) and LTE-Advanced are now seeing their first commercial infrastructure deployments.

1.2 MIMO Systems in Wireless Telecommunications

In the earliest wireless communication protocols, data transfer was achieved with a single transmitting antenna sending information to a single receiving antenna. This Single-Input Single-Output (SISO) system presented severe limitations in terms of inefficient use of spectrum, vulnerability to multipath and fading effects, and increased sensitivity to changes in Signal-to-Noise Ratio (SNR). The use of multiple antennas at the receiver lead to the development of Single-Input Multiple-Output (SIMO) wireless systems in which a variety of diversity schemes are available to help reduce the impact
of the weaknesses present in SISO systems. To further combat these problems and improve performance, the newest set of standards (802.11n, HSPA+, LTE, and their successors) utilize Multiple-Input Multiple-Output (MIMO) systems. MIMO communications systems use multiple antennas at the transmitter and receiver to improve capacity and protect against channel fading effects. While the earliest concepts in MIMO date back to the early- to mid-1970s [2, 3], it was not until the mid-1990s that more refined approaches to MIMO were considered, in which multiple antennas could be co-located at the transmitter and/or receiver to improve overall link throughput [4, 5]. With multiple antennas on both sides of the system, more than one communication link can be established between the transmitter and receiver (as opposed to the SISO case), and a bevy of tools such as precoding (beamforming), diversity coding, and spatial multiplexing can be used to improve overall data rates and compensate for a single link hindered by fading effects or destructive interference.

1.3 Motivation

In addition to key techniques such as beamforming and multiplexing, the physical antenna design can prominently affect the performance of MIMO systems. For example, existing MIMO solutions leverage pattern and polarization diversity offered by antennas to improve system performance [6]. Earlier theoretical work demonstrated room for capacity improvement in wireless channels using spatial multiplexing techniques [7, 8], however it appears that there is still room to fully investigate signal correlation and antenna coupling effects caused by keeping the antenna elements in close physical proximity [9] and alternative solutions for applications requiring closely-spaced antennas.

One generally accepted rule of thumb in MIMO antenna design specifies that radiating elements require approximately one half-wavelength separation to receive decorrelated
signals and independent fading paths [10]. This figure is often used to determine system parameters and allows for convenient design specifications of simple antenna arrays. However, there are many MIMO applications in which it is either desirable or required for radiating elements to be in close proximity with each other. Without proper design choices, the result is high signal correlation and mutual coupling between the antennas. This usually mars the system performance by causing poor antenna diversity and unwanted interaction between the antenna elements [11].

Previous related work has focused mainly on external hardware (via various matching and decorrelating networks) and coding methods to solve the issues that arise with closely-spaced antenna elements [12-16]. These methods are normally used in conjunction with a variety of strip antennas widely used in commercial applications, but one area that has not been thoroughly surveyed is the impact of physical antenna design on offsetting the coupling and correlation effects caused by small free-space antenna separation. Research in this area may provide a physical alternative to the existing hardware and software solutions for close-proximity spacing in MIMO antenna systems.

1.4 Objectives

The main goal of this thesis is to investigate the viability of a MIMO antenna design that uses closely-spaced radiating elements embedded in a high-permittivity dielectric material. The idea is to employ a high-permittivity dielectric in conjunction with a normal MIMO antenna array in order to provide the effect of larger element spacing (in wavelengths) than is physically present. The basic antenna design stems from the lens antenna family of aperture antennas, owing to their common construction using dielectric ceramics, high controllability of directivity, and simple integration with strip antenna feeds. A secondary effect of the lens design is improved pattern diversity
between the two radiating elements. In order to optimize the antenna design and evaluate its performance, further objectives need to be specified.

First, baselines for performance comparison need to be established to quantify the benefit that the antenna provides. Throughout this thesis, the common metric used to measure performance of the system is the channel capacity of a 2x2 (MIMO) system. The two benchmarks used for the assessment involve two strip dipoles in free-space, with separation based on the effective-wavelength separation of antennas in the lens, and the physical dimensions of the lens itself.

Next, the antenna design must be optimized to provide the best results. This is achieved through a series of parametric tests in which factors including material properties, dipole separation, and lens geometry are varied and compared.

Finally, the chosen antenna design must be put through more vigorous experimentation in order to validate its results from the design stages. For this purpose, a 2D environment more representative of a multipath-rich wireless channel is simulated, and the lens antenna is compared against the free-space benchmarks established earlier.

1.5 Thesis Outline

Chapter 2 contains a brief overview of the electromagnetics and wireless telecommunications concepts used to support the theory behind the design of the antenna. It also introduces the notion of channel capacity – the primary figure of merit used in the study. Finally, there is a review of some of the existing methods used to deal with closely-spaced antennas.

Chapter 3 describes the final antenna design and encompasses the process of design optimization through various parametric tests. An introduction is made to the
experimental setup used throughout the design process. Primarily, reasoning for design choices is provided alongside channel capacity comparisons and analysis of antenna patterns.

Chapter 4 demonstrates the results of extended testing of the chosen antenna design. This includes a variation to the original test setup, a 2D simulation of a wireless channel that more appropriately models a real-world environment, and the addition of new test metrics to further reinforce the results of the design process.

Chapter 5 summarizes the findings of the thesis, draws conclusions from test results, and remarks on further work that can be conducted in the research.
Chapter 2
Key Concepts in Wireless Telecommunications and Electromagnetics

This chapter serves as a primer for some of the principles and concepts used in the design of the antenna. A brief explanation of wireless systems is presented and the experiment’s main measure of performance – channel capacity – is introduced. Methods by which MIMO systems improve wireless data rates are reviewed, along with a summary of existing solutions to close-proximity antenna problems.

2.1 Wireless Systems Overview

As mentioned in Chapter 1, the use of MIMO systems in wireless telecommunications is a fairly recent development. Original wireless networks used a SISO system, in which a single transmitting antenna broadcasts to a single receiving antenna. While SISO systems are easier to implement since they do not require multiple-antenna coding solutions and have fewer components, they suffer the drawback of relying on a single link to transfer all information. Multiple effects such as channel fading and multipath propagation can cause the SNR at the receiver to drop to a point where data cannot be received and decoded reliably. The introduction of multiple-antenna systems introduces the potential for faster, more reliable transfer of information.

2.1.1 SIMO and MISO Systems

Once extra antennas are introduced at the receiving or transmitting end of the link, the system becomes a Single-Input Multiple-Output (SIMO) or Multiple-Input Single-Output (MISO) system, respectively. The addition of these antennas increases the
number of possible links between the transmitter and receiver, providing protection against multipath and fading effects as well as the opportunity to increase the rate of data transfer by utilizing a variety of diversity schemes [10].

SIMO setups improve system performance by using the received signal at each antenna to form the best possible recreation of the signal on the transmitting end. In well-characterized channels, this can be done using a form of maximal-ratio combining, adjusting the gain of each channel proportionally to the signal strength and summing the signals over the multiple receivers. In more primitive systems, the strongest received signal at any one time may be used and the other information discarded completely (selection combining).

MISO systems can also leverage various methods to take advantage of the additional informational links. For instance, in a characterized channel environment, multi-stream beamforming (precoding) gives the signal emitted from each transmitting antenna an appropriate phase weighting such that the signal power is maximized at the receiver. Alternatively, there is a selection of diversity coding techniques, in which a single signal is transmitted using full or near orthogonal space-time codes, allowing the receiver to fully reconstruct the original signal accurately. Diversity coding is useful in wireless systems with no channel knowledge, but does not benefit from beamforming or array gains.

2.1.2 MIMO Systems

In Multiple-Input Multiple-Output (MIMO), the system consists of multiple antennas on both the transmitting and receiving ends. This makes it possible to employ any of the performance-improving methods used in both SIMO and MISO systems. Moreover, MIMO systems can take advantage of spatial multiplexing, in which a high-rate signal can be split into multiple streams, each transmitted from a different transmit antenna.
If the receiving antennas can collect the signals with adequate spatial recognition, the streams can be received in parallel channels and reconstructed, resulting in a much higher throughput in the system. Thus, while diversity coding aims to increase system robustness, spatial multiplexing is designed to maximize data rate. The effectiveness of spatial multiplexing relies on two main factors: the number of streams supported (the maximum number of streams available is limited by whichever end of the system has less antennas) and the channel in which the wireless system is operating. Additionally, the efficacy of the technique can be improved by promoting system diversity through a variety of means such as antenna pattern diversity, use of orthogonal polarizations, and spatial diversity achieved with sufficient element separation. Consequently, in many MIMO systems that are designed to take advantage of spatial multiplexing, strong pattern diversity can play an important role.

Wireless communication systems are often described as being MxN, where M is the number of transmitters, and N is the number of receivers. Common MIMO configurations are 2x2, 3x3, and 4x4, with higher antenna densities possible. In this study, the system is tested using a 2x2 setup. The choice is representative of the most common MIMO systems, as most transceiver chipsets do not yet support the use of more than two spatial streams, and the results can be used to demonstrate advantages of the antenna design that may be carried over to MIMO systems of higher dimensions.

2.1.3 Channel Capacity

In order to assess the performance of a wireless system, a figure of merit must be chosen. One of the common measurements used in communications and information theory is channel capacity, $K$. The channel capacity of a system measures the upper bound of the system’s ability to transfer information reliably, and can be measured in bits/s/Hz. Channel capacity is determined by the environment in which the wireless
system operates and the SNR of the system. Thus, in order to calculate channel
capacity, the channel must be characterized either from previous knowledge or
empirical testing.

The characterization of the wireless channel can be described using the channel matrix,
\( \mathbf{H} \). In narrowband wireless systems, \( \mathbf{H} \) is an \( M \times N \) matrix of complex values \( h_{ij} \), for \( i = 1, \ldots, M \) and \( j = 1, \ldots, N \), where each value \( h_{ij} \) is the transfer function between
transmitting antenna \( j \) and receiving antenna \( i \). A general channel matrix can be
represented as

\[
\mathbf{H} = \begin{bmatrix}
h_{11} & \cdots & h_{1N} \\
\vdots & \ddots & \vdots \\
h_{M1} & \cdots & h_{MN}
\end{bmatrix}.
\] (2.1)

For a 2x2 MIMO system there are four transfer function values. In this study, the
values are determined by operating each of the four links separately with transmitted
signal strength of \( 1 \angle 0 \). The system response for the particular wireless link is then
obtained by summing the received multipath signals.

Once the channel matrix has been fully described, the response of the system can be
modelled simply by

\[
\mathbf{y} = \mathbf{Hx} + \mathbf{n},
\] (2.2)

where \( \mathbf{y} \) and \( \mathbf{x} \) are the receive and transmit vectors, respectively, and \( \mathbf{n} \) is the noise
vector [17]. Further, channel capacity can be calculated using Shannon’s Equation [18],

\[
K = \log_2 \det \left( \mathbf{I} + \frac{\rho}{N_t} \mathbf{HH}^H \right),
\] (2.3)

where \( \rho \) is the SNR and \( N_t \) is the number of transmitters. Since system noise can vary
due to a number of factors and it is plain that SNR affects channel capacity
calculations, $K$ is often calculated with a fixed value for SNR or it is plotted over a range. The main point of interest in channel capacity calculations of this study is the effect of physical antenna design on the capacity of the system. Varying the antenna design can affect antenna pattern diversity, signal correlation, and element coupling properties, which in turn will alter the transmission coefficients that form the channel matrix (given a consistent test environment). For this reason, transmit power control methods are ignored in the calculation.

### 2.2 Material Properties and Effective Wavelength

One of the current problems faced by MIMO technology is the space required to achieve low correlation and coupling between groups of transmitting and receiving antennas [19]. Antennas used in general MIMO studies (such as strip monopole and dipole antennas) normally do not perform well in close proximity with each other due to interference caused by mutual coupling and undesired parasitic effects. In addition, their small physical separation and omnidirectional antenna patterns cannot realize some of the advantages of MIMO systems, such as decorrelated fading channels and spatial diversity. As discussed earlier, many of these impediments have a much smaller impact once the antenna elements have a separation of roughly one half-wavelength, but it may not always be practical to provide that physical spacing.

The approach taken in this thesis is to embed the antennas in a high-permittivity dielectric, leveraging the material’s refractive index, $n$. The refractive index is related to the material properties of relative permittivity, $\varepsilon_r$, and relative permeability, $\mu_r$:

$$n = \sqrt{\varepsilon_r \mu_r}$$

(2.4)

In most naturally occurring materials, $\mu_r \cong 1$ at radio frequencies, and $n$ can be approximated as $\sqrt{\varepsilon_r}$. In the dielectric, the wavelength at a specific frequency is then
reduced from the free-space wavelength by a factor of $n$, resulting in the effective wavelength:

$$\lambda_{\text{eff}} = \frac{\lambda_f}{n}$$ (2.5)

With the reduced effective wavelength within the dielectric, the required physical separation is reduced to $\lambda_{\text{eff}}/2$, allowing for more compact antenna placement.

However, for this approach to work, the dielectric must fill the entire space of the system, which is not possible. Otherwise, interactions at the dielectric-air boundary will result in detrimental effects like wave refraction, internal reflections, and signal dispersion. These effects may be mitigated or minimized by choosing an appropriate antenna design, as shown in the next chapter.

2.3 Alternative Solutions for Closely-Spaced Antennas

Other work in the area of MIMO antenna systems has resulted in various solutions to the coupling and correlation problems presented by closely-spaced antenna arrangements, as well as studies of the actual impact on channel capacity caused by these issues [20]. Before proceeding with the design and assessment of a new approach, it is important to understand the advantages and drawbacks of existing methods.

2.3.1 Decorrelating Networks

One of the most common solutions to problems involving poor signal correlation is the use of decorrelating networks. The network is a physical addition to the antenna system and improves signal decorrelation between two or more antennas, depending on its design. A properly designed network benefits from being able to decorrelate signals independent of the element type or spacing, reducing envelope correlation and
improving channel capacity [12]. The major drawback of decorrelating networks is the additional physical space required to implement the network.

2.3.2 Precoding

As described earlier in this chapter, one technique used to improve SNR in a multi-antenna system is beamforming, in which the same signal transmitted from each antenna is weighted and phase-shifted so that the resulting total signal at the receiver benefits from constructive interference of the multiple signals, maximizing the signal strength. Precoding is a generalization of beamforming that aims to maximize link throughput at the receiver by having the transmitter emit multiple data streams with independent weightings. Precoding allows for more robust transmission of information and enables multi-user MIMO systems to be served from a single transmitting array, compensating for poor signal correlation by using orthogonal coding methods to send multiple streams of data simultaneously, so it does not need to rely on improving signal decorrelation to achieve spatial diversity like other MIMO systems. However, in order to select appropriate precoding strategies, multiplexing schemes, and codebooks to maximize channel throughput and capacity, there must be significant or full knowledge of the channel state information ahead of implementation [17].

2.3.3 Isolating Fences

One technique used to reduce mutual coupling effects between closely-spaced antennas is to introduce a physical barrier. A simple perfect electric conductor (PEC) can be used as an isolating fence, and the design can be improved with the aid of more complex materials and structures, for example by using a metamaterial spacer acting as an artificial magnetic conductor (AMC) [21]. These barriers provide a solution that can improve channel capacity, create improved antenna pattern diversity, and effectively decorrelate and decouple closely-spaced antennas. Some of the drawbacks of using
spacers are the requirement of an additional physical component (possibly of high complexity if using an AMC structure) and the limited use in systems with more than two antennas, since the barriers effectively make the elements sectorized components.

2.3.4 Antenna Selection Methods

By operating a subset of available antennas at any given time, overall system performance can be improved. There are a variety of antenna selection methods, ranging from ‘hard’ selection, in which a predetermined set of antennas are active, to ‘soft’ selection methods in which all antennas are used in varying combinations. The best results are achieved with soft antenna selection, with continuing work on improved selection schemes being developed. Antenna selection is a good way to aid against highly correlated signals between antenna elements in MIMO systems, and can also achieve better spectral efficiency despite mutual coupling between elements [22]. Additionally, this is a useful method since it doesn’t require physical modifications to the system and can be effective in most MIMO applications. The limitations of this technique are mainly caused by high sensitivity to interelement spacing and no way to interact with the mutual coupling effects, making it better suited for MIMO systems that don’t require closely-spaced antenna elements, although there is still some recorded benefit in such applications.
Chapter 3
Design, Evaluation, and Parametric Study of the High-Permittivity Dielectric Lens Antenna

This chapter starts with an overview of the final antenna design presented in the thesis. The simulation setup used for design assessment and comparisons is expounded fully, followed by a thorough review of the design and decision-making process. Insight is provided by examining graphs of channel capacity for the various design candidates and elucidated with analysis of their antenna patterns.

3.1 Introduction to the Antenna Design

The antenna tested in this thesis is a dielectric lens antenna. Lenses are an appropriate choice since they conventionally use dielectric materials in their construction and can remain reasonably sized (e.g. lens diameters between $\lambda/2$ and $\lambda$). Lenses enable and smooth the transition of electromagnetic waves at the dielectric-air interface, and can potentially improve antenna directivity with a high degree of controllability. The chosen lens design is a stepped-index hemispherical lens, consisting of a high-permittivity core for the antenna feed and a lower permittivity matching layer used to reduce the effect of reflections at the lens-air interface [23].

The lens is designed for a two-element MIMO transmitter or receiver operating at 2.4 GHz. Figure 3.1 shows a top-down view of the lens and provides some of the antenna’s physical specifications. Figure 3.2 provides an isometric view of the lens, with the two lens layers shown in white, the two radiating elements in red, and the antennas’ feed points in green. The first layer of the lens is a hemisphere with a radius equal to a
Figure 3.1: Top-down schematic of the lens antenna design.

Figure 3.2: An isometric view of the lens antenna showing the two hemispherical layers, strip antenna placement, and antenna feed points. The red and green arrows correspond to the x- and y-axes in the top-down schematic.
quarter-wavelength in free-space, constructed with a dielectric of relative permittivity $\varepsilon_r = 16$. A second layer acts as a matching cap of thickness $\lambda_d/4$ and $\varepsilon_r = 4$, where $\lambda_d$ is the wavelength in the dielectric material. A hemispherical base lens was chosen due to its well-documented use in other areas of wireless communication – namely RADAR and satellite applications. The hemispherical design also allows for an easy to execute parametric design method with a few fixed and many variable properties. The geometry also makes the change in individual parameters more predictable compared to more complex lens structures such as ultrathin and flat lenses. The two radiating elements are centre-fed half-wave strip dipoles centered on the back of the lens with $\lambda_{lens}/2$ separation. Since the dipoles are located at a dielectric-air boundary, the effective wavelength of the dielectric half-space, $\lambda_{lens}$, must be used to calculate antenna separation:

$$\lambda_{lens} = \frac{\lambda_{fs}}{\sqrt{(1 + \varepsilon_r)/2}}$$  \hspace{1cm} (3.1)

Table 3.1: Final Design Specifications

<table>
<thead>
<tr>
<th>Component</th>
<th>Property</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Lens (Inner Layer)</td>
<td>Radius</td>
<td>31.25[mm]</td>
</tr>
<tr>
<td></td>
<td>Material</td>
<td>Dielectric ($\varepsilon_r = 16$)</td>
</tr>
<tr>
<td>Lens (Outer Layer)</td>
<td>Radius</td>
<td>46.875[mm]</td>
</tr>
<tr>
<td></td>
<td>Material</td>
<td>Dielectric ($\varepsilon_r = 4$)</td>
</tr>
<tr>
<td>Electrical Sources</td>
<td>Amplitude</td>
<td>1[V]</td>
</tr>
<tr>
<td></td>
<td>Impedance</td>
<td>50[Ω]</td>
</tr>
</tbody>
</table>
Table 3.1 shows the complete physical specifications of the antenna. The design was rendered and simulated using SEMCAD X – a 3-D full-wave electromagnetic simulation platform based on the finite-difference time-domain (FDTD) method – and post-processing was completed with MATLAB. The antenna was simulated using both harmonic and broadband simulations, with simulation details outlined in Table 3.2.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Simulation Mode</td>
<td>Harmonic/Broadband</td>
</tr>
<tr>
<td>Frequency</td>
<td>2.4[GHz]</td>
</tr>
<tr>
<td>Broadband Simulation Bandwidth</td>
<td>0.8[GHz]</td>
</tr>
<tr>
<td>Simulation Time</td>
<td>60 Periods</td>
</tr>
</tbody>
</table>

The results of the testing in this chapter revealed that the final lens antenna design outperformed the chosen test benchmarks and other candidate antennas, measured through the channel capacity of the antenna in a 2 x 2 MIMO environment. Further, exploration into several design variations will reveal that there is no strong correlation between antenna directivity and channel capacity, as was assumed early in the design process. The coupling characteristics of the radiating elements proved to be similar to a two-element free-space array (covered in more detail in Chapter 4), allowing for the
hypothesis that the improvement in capacity is mainly due to the lens antenna’s improved pattern diversity.

3.2 The Ring of Scatterers Channel Model

In the design of the lens antenna, performance was evaluated using a MIMO channel capacity simulation employing a channel consisting of a ring of isotropic scatterers. This Ring of Scatterers (RoS) model is used in various simulations to study the effects of Rayleigh fading paths and is a well documented model for propagation and interference in multipath-rich environments [24, 25]. In addition to exploring the effects of various parameter changes on the dielectric lens’ performance, the simulation focused the design on capacity improvement via the antenna’s pattern diversity. Further MIMO factors such as spatial multiplexing were accounted for in a full-wave simulation that will be presented in Chapter 4.

The following figures show the layout and geometry of the RoS model. As seen in Figure 3.3, the lens antenna sits at the middle of a ring of scattering elements, with two

![Figure 3.3: The Ring of Scatterers test setup and examples of multipath propagation routes.](image)
transmitting antennas (strip dipole antennas in free-space) some distance in front of the lens. Each transmitter is operated independently, causing the signal at the receiving antenna to be the superposition of the line of sight (LoS) signal and the various multipath signals from each scatterer. In Figure 3.3, the LoS path is shown as the bolded black line between the top transmitting antenna and the receiver, while examples of multipath reflections from scattering elements in front of and behind the lens are shown in blue and green, respectively. Two baseline cases are used for comparison against the lens antenna: two half-wave dipoles in free-space with separation equal to $\lambda/2$ and $3\lambda/4$. The first case acts as a free-space comparison to the equivalent separation of the radiating elements in dielectric wavelengths and the second case is a comparison to the physical size (diameter) of the final lens design.

Figure 3.4 shows the basis for the geometry required to calculate the total received signal. If the distance from the receiver to the center of the two transmitting antennas is $d_I$ and the separation between the two transmitting antennas is $s$, the LoS distance from the transmitter to receiver, $d$, can be expressed as

![Figure 3.4: Geometry of the Ring of Scatterers model.](image-url)
\[
d = \sqrt{d_1^2 + (s/2)^2}. \tag{3.2}
\]

arriving at an angle of
\[
\phi_{\text{LoS}} = \cos^{-1}(d_1/d). \tag{3.3}
\]

For each multipath signal, the total path distance is the sum of the scattering ring radius, \( r \), and the distance from the transmitter to scatterer \( n \), \( x_n \), which can be calculated using cosine law with knowledge of the angle of arrival for scatterer \( n \), \( \phi_n \). Thus, for each multipath signal, the total path length, \( l_n \), can be calculated as
\[
l_n = r + \sqrt{r^2 + d^2 - 2rd \cos(\phi_n - \phi_{\text{LoS}})}. \tag{3.4}
\]

Now, if the transmitting antenna broadcasts a signal of strength \( 1 \angle 0 \), LoS distance \( d \) along with phase constant \( \beta = 2\pi/\lambda \) can be used to determine the phase of the arriving LoS signal. The path length of each scattered signal \( l_n \) can likewise be used to determine the multipath signals’ phase offset, while amplitude modifications are determined by individual scatterer weights. Finally, each signal is modified depending on its angle of arrival using E-field and phase data from the simulated antenna pattern and then summed to obtain the total received signal for the transmitter-receiver link. Mathematically, the channel matrix coefficients \( h_{ij} \) can be expressed as
\[
h_{ij} = (10^{E(\phi_{\text{LoS}})/10}e^{i(d\beta+P(\phi_{\text{LoS}}))}) + \sum_{k=1}^{n} \Gamma_k(10^{E(\phi_k)/10}e^{i(l_k\beta+P(\phi_k))}) \tag{3.5}
\]

where \( E(\phi) \) and \( P(\phi) \) are the antenna pattern E-field and phase modifiers (in dB and radians, respectively) at angle \( \phi \) for receiving antenna \( i \). The first part of the equation
calculates the strength and phase of the LoS signal, while the summation term does the same for each of the \( n \) scattering elements. Free space path loss (FSPL) is omitted from this calculation since the effect of the term, as demonstrated later in this section, does not vary significantly between multipath signals. The result is that FSPL can be represented by a single scaling term, which does not affect capacity calculations and is thus discarded from the channel matrix equation.

The process is repeated for each of the four link combinations in a 2x2 MIMO system, resulting in a fully characterized channel matrix \( H \) that can then be used to calculate channel capacity \( K \) using Equation 2.3. In order to remove the element of SNR gain from the channel matrix, it was normalized by dividing each element, \( h_{ij} \), by the Frobenius norm of the matrix, calculated for an \( m \times n \) matrix as follows:

\[
\|H\|_F = \sqrt{\sum_{i=1}^m \sum_{j=1}^n |h_{ij}|^2}
\]

\[
H_{\text{norm}} = \frac{H}{\|H\|_F}
\]

The parameters used in the MIMO RoS simulation are outlined in Table 3.3. The number of scattering elements was chosen to be within the range normally used for one-ring scattering models (20-200) [26, 27] and distance parameters were chosen to be sufficiently far apart to put the antenna arrays in the far-field and provide noticeable phase change in the incident multipath signals. The 59 scattering elements (plus LoS signal) are treated as evenly distributed around the scatterer ring, resulting in an angular separation of six degrees between scatterers, and are fixed with uniform weighting and no additional phase modification (for this portion of the study, all scatterers are treated as PECs with reflection coefficient \( \Gamma = -1 \)).
Table 3.3: Ring of Scatterers Test Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>n</td>
<td>Number of scattering elements</td>
<td>59</td>
</tr>
<tr>
<td>$\Gamma$</td>
<td>Scatterer reflection coefficient</td>
<td>-1</td>
</tr>
<tr>
<td>$r$</td>
<td>Scatterer ring radius</td>
<td>50[m]</td>
</tr>
<tr>
<td>$d_1$</td>
<td>Transmitter-to-Receiver distance</td>
<td>1000[m]</td>
</tr>
<tr>
<td>$s$</td>
<td>Transmitter dipole separation</td>
<td>62.5[mm]</td>
</tr>
</tbody>
</table>

For sufficiently large $d_1$ and small $s$ as in the test parameters, $d \approx d_1$, and the longest scattering path will occur when the scatterer is on the opposite side of the lens so that $l \approx d + 2r$. The difference in path length between the LoS and scattering signals causes phase shifts at the receiver, but may also affect the received signal amplitude due to free space path loss which can be calculated using:

$$FSPL = \left(\frac{4\pi d}{\lambda}\right)^2$$  \hspace{1cm} (3.8)

Given the FSPL calculation in Equation 3.8, the maximum difference in path-loss between the LoS signal and a multipath signal (in dB) can be approximated as follows given knowledge of the operating wavelength (0.125 m) and test parameters from Table 3.3:

$$\Delta FSPL_{\text{max}} \approx 10 \log_{10}\left(\left(\frac{4\pi l_{\text{LoS}}}{\lambda}\right)^2 / \left(\frac{4\pi d}{\lambda}\right)^2\right) \approx 0.828$$  \hspace{1cm} (3.9)
Given this result, FSPL was not taken into account in the RoS simulation since the transmitter-to-receiver distance and the radius of the scattering ring were chosen so that the different multipath signals do not produce significant variations in FSPL.

### 3.3 Parametric Study

The optimization of the antenna design was accomplished by carrying out series of simulations in which various parameters such as element spacing, lens size, and material properties were isolated and adjusted. The most promising candidates were then further refined and additional effects such as dielectric losses, directivity, and internal lens reflections were considered.
3.3.1 Establishing Benchmarks

At the beginning of the design process, the test used for comparison of designs was the performance of two strip dipoles with equivalent free-space separation of radiating elements. Early lens designs experimented with both half- and full-effective wavelength separation of antennas, and the competing designs were weighed against their respective free-space results. Once a more definitive lens design had been chosen, the $3\lambda/4$ free-space separation case was added to compare against the physical size of the lens. Figure 3.5 shows a graph of the capacity performance versus SNR for each of the benchmark cases and full-wavelength separation (dashed lines), as well as several smaller dipole separations in free-space (solid lines). The capacity is graphed against receive SNR between -10 and 20 dB. Receive SNR was chosen in order to normalize the effect of capacity improvements due to SNR gain from the dielectric lens compared to

![Figure 3.6: Capacity at 0 dB receive SNR versus equivalent element separation in free-space ($\lambda_{fs}$) and lens ($\lambda_{lens}$) wavelengths.](image)
the antenna arrays in free-space. Early tests were not normalized, effectively comparing
capacity to transmit SNR and resulting in more favourably weighted results for the
dielectric lens antennas where the increase in capacity could be attributed to factors
other than better signal decorrelation or antenna pattern diversity (i.e. SNR gain).
These results were discarded in favour of the normalized receive SNR results.

Figure 3.6 shows a post-design evaluation of the effect of the lens at various element
separations. At very small element spacing (1/10- and 1/8-wavelength separation), the
lens succeeded in making the effective spacing in terms of lens wavelengths (as defined

![Normalized H-plane patterns of $E_\phi$ for two active dipoles with $\lambda/2$
separation in free-space, $\varepsilon_r=16$ dielectric, $\varepsilon_r=16$ sphere with $\varepsilon_r=4$ matching layer, and
the final lens design.](image)

Figure 3.7: Normalized H-plane patterns of $E_\phi$ for two active dipoles with $\lambda/2$
separation in free-space, $\varepsilon_r=16$ dielectric, $\varepsilon_r=16$ sphere with $\varepsilon_r=4$ matching layer, and
the final lens design.)
in Equation 3.1) look similar to the equivalent wavelength-separation of elements in free-space. At larger separations (1/4- and 1/2-wavelength) there is a noticeable improvement in channel capacity. This may be attributed to improved pattern diversity visible from the lens’ antenna pattern, as will be discussed later.

Besides using the RoS model to provide numerical comparisons, additional information could also be gleaned by examining the antenna patterns of the candidate designs. Observing the antenna pattern produced when both radiating elements are active offers an experiential benchmark for comparisons with the lens designs. Meanwhile, looking at the antenna pattern when only one of the dipoles is active can help in identifying designs where there is sufficient pattern diversity to increase channel capacity through spatial multiplexing. Figure 3.7 shows the normalized antenna patterns of the $\lambda/2$-separation case alongside the patterns produced by two dipoles with $\lambda_d/2$-separation in an infinite space of $\varepsilon_r = 16$, a $\varepsilon_r = 16$ spherical lens with $\varepsilon_r = 4$ matching layer, and the antenna pattern of the final lens design (with $\lambda_{lens}/2$-separation on a hemispherical lens antenna). Notably, it can be observed that the patterns of the lens and the completely dielectric space are very similar in shape to the free-space case, suggesting that in the dielectric the smaller physical separation behaves as expected with reduced dielectric wavelength.

This also confirms that the lens successfully preserves the antenna pattern as it is carried into the far-field from the dielectric, emulating the pattern of two dipoles with half-wavelength-separation in free-space. From this point, the lens can provide additional pattern diversity between the radiating elements, resulting in improvements in channel capacity.
3.3.2 Material Selection and Element Spacing

The initial stages of design cover a broad range of options so that the most promising results can be selected for further study. In the first set of tests, both half- and full-effective wavelength separations of dipoles were investigated. A hemispherical lens of radius \( r = \lambda/4 \) commonly used in basic lens antenna design was the basis for comparisons to the free-space benchmarks. The materials examined in the test were mainly magnetically matched materials (\( \varepsilon_r = \mu_r \)) with relative permittivity and permeability ranging from four to ten, as well as a non-magnetic dielectric of material properties \( \varepsilon_r = 16, \mu_r = 1 \). Figures 3.8 and 3.9 show the channel capacity graphs for the ull- and half-wavelength antenna separations, respectively. In Figure 3.8, it can be observed that none of the lenses outperform the baseline case of two dipoles in free-space with one wavelength separation (and consequently lenses with full-wavelength element separation should not be pursued further). It is possible that this is the case since the two-antenna array in free-space achieves significantly low signal correlation between radiating elements due to the relatively large inter-element spacing. In addition, due to the limited size of the lens, using full-effective-wavelength separation – especially in the case of materials with lower dielectric constant and therefore larger \( \lambda_{\text{lens}} \) length – may place the antennas too close to the edge of the dielectric for the lens to propagate the antenna pattern as intended. This is a possible explanation for why the matched lenses with lower relative permeability (i.e. \( \varepsilon_r = 4, 6 \)) were the worst-performing materials. In Figure 3.9 all of the lenses show improvements in channel capacity compared to the baseline (and so are all possible candidates for additional assessment).
Figure 3.8: Capacity versus SNR of hemispherical lenses with varying material properties and $\lambda_{\text{lens}}$ element spacing. $\lambda/2$ lens diameter.

Figure 3.9: Capacity versus SNR of hemispherical lenses with varying material properties and $\lambda_{\text{lens}}/2$ element spacing. $\lambda/2$ lens diameter.
Figure 3.10: Capacity versus SNR of spherical lenses with varying material properties and $\lambda_d/2$ element spacing. $\lambda/2$ lens diameter.

Figure 3.11: Capacity versus SNR of hemispherical and spherical lenses for the three best-performing material property combinations. $\lambda/2$ lens diameter.
This suggests that the closer spacing places the radiating elements sufficiently far from the edge of the lens to avoid the undesired behaviour seen in the tests with full-wavelength separation, and the improvement in capacity in fact suggests that the lens is providing some combination of improved signal decorrelation, decoupling, antenna pattern diversity, or directivity improvement. One area of concern in this line of investigation is that the capacity improvement may be due solely to the increased directivity of the hemispherical lens instead of an improvement in pattern diversity. To test this theory, channel capacity was measured for fully spherical lenses in which two dipoles are embedded in the center of a sphere with half-dielectric wavelength separation. Figure 3.10 shows the channel capacity for the spherical lenses.

At this point, the most effective material properties were identified for further experiments. The chosen lenses had $\varepsilon_r = \mu_r = 8$, $\varepsilon_r = \mu_r = 10$, and $\varepsilon_r = 16, \mu_r = 1$.

Figure 3.11 shows the channel capacity versus SNR graph for the three configurations comparing the hemispherical and spherical lenses. It can be seen that there is a significant improvement in capacity between the two designs only in the $\varepsilon_r = \mu_r = 10$. The $\varepsilon_r = \mu_r = 8$ improvement is marginal, and there is no discernible difference in the $\varepsilon_r = 16, \mu_r = 1$ lenses. This suggests that either the shape of the lens in combination with the material choice does not provide a significant improvement in directivity of the lens, or that in the case of this lens design improvements in capacity are due mainly to better pattern diversity and not strongly correlated to the directivity of the lens.

### 3.3.3 Diameter Sweep

The next set of tests alter the lens diameter in order to see if the size of the lens can be reduced while maintaining the improved channel capacity of the original $\lambda/4$-radius lens and explore the effects of varying the lens size with the matched materials. Lens size was set to 90%, 95%, 105%, and 110% of the original specification for all lenses, and
Figure 3.12: Capacity versus SNR of hemispherical lenses with $\varepsilon_r=\mu_r=8$ and varying lens diameters $D$.

Figure 3.13: Capacity versus SNR of hemispherical lenses with $\varepsilon_r=\mu_r=10$ varying lens diameters $D$. 
the matched lenses were also tested at 200% and 300% lens diameters. Figures 3.12, 3.13, and 3.14 show the results of the variations for the $\varepsilon_r=16$, $\mu_r=1$ and varying lens diameters $D$.

In all three cases, varying the lens radius around $\lambda/4$ resulted in a noticeable reduction in channel capacity in the range of 15-25%. This effect may have been caused by the closer proximity of the strip elements to the edges of the lens (and the dielectric-air boundary) since placing the edges of the antennas close to the lens edge no longer simulates fully embedding the antennas in the dielectric material, making the resulting antenna patterns unpredictable. Additionally, increasing the lens size to twice and three times its original value in the matched cases showed a small increase in capacity in the 5-10% range. This improvement was not enough to justify doubling or tripling the size of the lens, as the volume of space occupied by the lens is proportional to the cube of the radius, and the benefit of increasing the radius by that magnitude can quickly be
surpassed by more compact antennas. With these results, it is sensible for the lens to be kept at its original size.

### 3.3.4 Effects of Dielectric Losses

Up to this point, the various lenses were simulated and channel capacity was calculated taking into account only properties produced by varying physical parameters in a lossless environment. With the selection of a lens shape, size, and material choices that provide a significant channel capacity improvement, it is necessary to explore the impact of dielectric losses that would be present if the lens were fabricated for real world use. The channel capacity was recalculated for loss tangents of 0.001 on the low

![Figure 3.15: Capacity versus SNR of hemispherical lenses with dielectric losses. Solid lines show the lossless case, dashed lines tan δ=0.001, and dotted lines tan δ=0.01.](image)
end and 0.01 on the high end. The results, shown in Figure 3.15, demonstrate that even small dielectric losses greatly impact the channel capacity of all three basic lens configurations but have a far lesser impact on the matched lenses of increased size. To better understand what is causing the degradation, a heat map of electric field strength can be viewed at a slice through the center of the lens. Figure 3.16 shows such visualizations for the three candidate lenses. As shown earlier in Figures 3.12 and 3.13, the matched material lenses showed slight degradation in channel capacity when the size of the lens was varied around a radius of $\lambda/4$, but had an improved capacity when a larger lens size was used. This may indicate that the $\lambda/4$-radius lenses do not behave as though the radiating elements are embedded properly in the matched materials. In Figure 3.16, the two matched lenses appear to be operating with multiple internal reflections – an unexpected property that could possibly be related to the size of the lens and account for capacity deterioration. This is reinforced by capacity results of
the $\varepsilon_r = \mu_r = 8$ and $\varepsilon_r = \mu_r = 10$ lenses with larger radii. In Figure 3.15 the larger variations of the matched material lenses showed much smaller drops in capacity when dielectric losses were taken into account, suggesting that there is a minimum threshold of lens size required before the matched materials will behave as expected. In the case of the $\varepsilon_r = 16, \mu = 1$ lens, there are only minor internal reflections, but the mismatch between the dielectric material and the air at the lens boundary results in a large amount of energy lost, explaining the decline in channel capacity in the purely dielectric case. The problem of mismatches causing losses at material boundaries is not uncommon in lens antenna design, and as shown in the following section there are design choices that can be applied to alleviate the drawback. However, the issue with the matched lenses appears to be related to the lens size, not the material properties, so the matched materials may not be an appropriate choice for an antenna of this nature. Hence, for the remainder of the study the design will focus on purely non-magnetic ($\mu_r = 1$) dielectric lenses.

3.3.5 Stepped-Index Matching Layer

In order to reduce the effect of the material mismatch at the dielectric-air boundary for the $\varepsilon_r = 16$ lens, a matching layer was added to the lens design, forming a stepped-index lens. Gradient-index lenses, in which the material of the lens contains a gradual variation in refractive index, are commonly used in optics and electromagnetics to produce lenses with flat surfaces or without aberrations that may otherwise be unavoidable due to the lens’ shape or material properties [28]. A stepped-index lens is a gradient-index lens in which the variations in material properties are discrete, making fabrication processes simpler. Such lenses are used commonly in electromagnetic communications and can allow for a smoother passage of electromagnetic signals between materials of high refractive index and free-space. In the implementation of the
stepped-index matching layer, the improvement in capacity versus the increase in lens size was considered, along with the new results for capacity in which dielectric losses were taken into account. The effect of varying the permittivity of the dielectric was then explored, and adjusting the option of adjusting lens size was revisited.

3.3.5.1 Introduction of the Matching Cap

The matching layer used to form the stepped-index lens is a hemispherical cap that has relative permittivity and thickness defined by Equations 3.10 and 3.11, respectively:

\[
\varepsilon_{r-step} = \sqrt{\varepsilon_r} = 4 \tag{3.10}
\]

\[
t = \lambda_{d-step}/4 = 0.0156[m] \tag{3.11}
\]

Since the matching cap encompasses the entire hemispherical lens, it increases the diameter of the lens to \(3\lambda/4\). With this dimension in mind, a second benchmark for channel capacity is established; in addition to being compared to a two-antenna array with \(\lambda/2\) separation (equivalent element spacing), the results are also compared to a similar array with \(3\lambda/4\) separation (equivalent physical dimensions). Figure 3.17 shows the channel capacity graph for the \(\varepsilon_r = 16, \mu_r = 1\) hemispherical lens with and without the \(\varepsilon_r = 4, \mu_r = 1\) matching layer. Not only does the matching layer significantly increase the channel capacity of the system, it also provides similar capacity results when dielectric losses are added to the simulation.
Since the \(\varepsilon_r = 16\) lens was the only non-matched lens explored in the original set of material tests, other relative permittivity values should be explored. This can be used to determine if there is an optimal range where the channel capacity of the lens is maximized, or if the physical element separation can further be reduced by increasing the refractive index of the lens material. To do this, the \(\lambda/4\)-radius lens with matching layer was tested over a range of material properties in which only the relative permittivity was varied, similar to the original matched material test.

In order to choose an appropriate range of dielectric constants to examine, some design considerations should be contemplated. If the material has a lower relative permittivity,
the required physical element spacing becomes larger and may put the edges of the antennas too close to the edge of the lens for it to behave as expected. On the other end, dielectric materials with very high permittivity ($\varepsilon_r \geq 50$) tend to consist of ceramic-polymer composites that may be brittle and have low dielectric and mechanical strength [29]. Commonly, single-antenna stepped-index lens antennas used for applications such as satellite communications may use dielectrics materials with relative permittivity as high as $\varepsilon_r = 30$. It is also reasonable to assume that at some point (with a very high relative permittivity) the dielectric wavelength will be so small relative to the size of the strip antenna that the physical element spacing will not provide sufficient electrical decoupling to simulate the desired equivalent free-space separation. With these details in mind, a range of $12 \leq \varepsilon_r \leq 25$ was chosen as the basis for the relative permittivity sweep, with the opportunity to expand to higher values if results were promising. The test points chosen and the accompanying physical element spacing and matching layer properties are outlined in Table 3.4.

Table 3.4: Relative Permittivity Sweep Test Parameters

<table>
<thead>
<tr>
<th>Lens $\varepsilon_r$</th>
<th>Dipole Separation</th>
<th>Matching Layer $\varepsilon_{r\text{-step}}$</th>
<th>Matching Layer Thickness</th>
</tr>
</thead>
<tbody>
<tr>
<td>12 0.0245[m]</td>
<td>3.464</td>
<td>0.0180[m]</td>
<td></td>
</tr>
<tr>
<td>16 0.0214[m]</td>
<td>4</td>
<td>0.0156[m]</td>
<td></td>
</tr>
<tr>
<td>20 0.0193[m]</td>
<td>4.472</td>
<td>0.0140[m]</td>
<td></td>
</tr>
<tr>
<td>25 0.0173[m]</td>
<td>5</td>
<td>0.0125[m]</td>
<td></td>
</tr>
</tbody>
</table>
The results of the various materials tested can be seen in Figure 3.18. It is apparent that all of the lenses performed better than both of the benchmark test cases, but the original $\varepsilon_r = 16$ lens with matching layer had the best results. This suggests that there is in fact a ‘sweet spot’ beyond which increasing the relative permittivity does not provide further improvements in capacity. It should also be noted that lenses with higher relative permittivity are smaller since their matching layers are less thick as shown in Table 3.4. However, the difference is marginal compared to the difference in channel capacity since the lens radius excluding matching layer is constant.

### 3.3.5.3 Radius Sweep

With the results supporting the $\varepsilon_r = 16$ stepped-index lens design, adjusting the size of the lens can once again be revisited. Since the matching layer remains a fixed thickness relative to its material, the radius of the $\varepsilon_r = 16$ layer must be adjusted. The goal at

![Figure 3.18: Capacity versus SNR of stepped-index hemispherical lenses with varying dielectric constants. $3\lambda/4$ lens diameter.](image)
this point is to make the high-permittivity layer smaller to compensate for the additional physical space occupied by the matching layer.

The $\varepsilon_r = 16$ layer radius was tested at 90% and 80% of the original $\lambda/4$ radius. Combined with the matching layer, this resulted in total lens diameters of $0.75\lambda$ (original lens size), $0.7\lambda$, and $0.65\lambda$. The results are shown in Figure 3.19, and it is clear that reducing the inner lens radius has a significant impact on the channel capacity of the lens – both of the smaller lens sizes resulted in a channel capacity approximately 25% lower than the original lens size. Although they both still slightly outperformed the two-antenna array with $3\lambda/4$ separation, the capacity shortfall cannot justify the reduction in lens size.

Figure 3.19: Capacity versus SNR of the stepped-index hemispherical lens with varying $\varepsilon_r=16$ layer radius.
3.3.6 Lens Extensions and Effects of Directivity

With the material of the lens chosen and the addition of the matching layer, the effect of lens directivity on channel capacity can be revisited in more detail. To determine if there was any correlation between directivity and improved capacity, variations of the lens design were considered. As in the earlier material selection tests, a full spherical lens (now with matching layer) was considered. From examination of Figure 3.7, the antenna pattern was identical to the pattern produced by two antennas with $\lambda_d/2$ separation in a full-dielectric space, with very similar front lobe shape to the pattern produced by the hemispherical lens with matching layer. Given the small back lobe of the hemispherical lens, it can be postulated that the hemispherical lens with matching layer has at most an additional 3 dB of gain in directivity compared to the equivalent full spherical lens. Thus, by comparing the channel capacity calculations for these two lenses, a strong argument can be made for or against the positive effect of directivity on the channel capacity of the lens design. However, to be thorough in the investigation

![Figure 3.20: Geometry of extended hemispherical lenses.](image)
extended hemispherical lenses (which generally have even better directivity properties) were also contemplated. An extended hemispherical lens is a lens consisting of a hemispherical component with the addition of a cylindrical extension of the same material and radius. The geometry the lens is shown in Figure 3.20. Depending on the length of the extension, \( l \), the result is a lens design with better robustness against spherical/comatic aberrations (important in optical applications) and higher directivity (useful in electromagnetic applications). Two types of extended hemispherical lenses were tested: the hyperhemispherical lens and an empirically defined extended lens. A hyperhemispherical lens is an extended lens designed to be aplanatic, or free from aberrations normally caused by the lens curvature. The lens length, \( d \), is a function of the lens’ index of refraction and radius:

\[
d = r \left( \frac{n + 1}{n} \right)
\]  

(3.12)

Given that the radius of the lens with matching layer is fixed at \( 3\lambda/4 \) and we know \( n \) to be 4 for the \( \varepsilon_r = 16 \) lens, \( l \), can be calculated:

\[
l = r \left( \frac{n + 1}{n} - 1 \right) = \frac{3\lambda}{16}
\]  

(3.13)

The second extended hemispherical lens used to test the effects of directivity is based on a study in which optimal extension lengths were determined experimentally [30].
In the study, it was determined that for a single double-slot antenna mounted on an extended hemispherical lens with λ/4 matching layer, an extension length of 32-35% of the lens radius provided reduced interelement signal coupling and improved lens directivity. Similarly, for multi-element imaging applications, an extension length of 38-39% of the lens radius provided the best results. Both extension ranges were tested and the capacity results for all the lens variations can be seen in Figure 3.21. It is notable that even though the extended hemispherical lenses have higher demonstrated directivity (in the range of 7.005-7.755 dBi), none of them outperformed the basic hemispherical lens with matching layer (which had a peak directivity of 6.813 dBi). Additionally, the spherical lens with matching layer showed a significant improvement in capacity compared to the equivalent two-antenna array in free-space. The poorer
performance of the extended lenses suggests that there is no direct correlation between lens directivity and channel capacity for these lens applications. Further, the comparison between the hemispherical and spherical lenses and the free-space benchmarks suggest that the high-permittivity lens with matching layer provides an improvement in the pattern diversity between the two radiating elements that consequently results in better signal decorrelation and improved channel capacity.

### 3.3.7 Factors Contributing to Capacity Improvement

The iterative design and testing process resulted in a lens antenna design that shows marked improvements in channel capacity over comparable two-element arrays in free-space. While the next chapter will focus on more rigorous testing and verification of the design as well as more complete characterization of the antenna properties, it is important to look at an aspect of design that may contribute to the improved capacity calculations. If the lens antenna successfully makes the electrical spacing of the two dipoles appear larger than the physical spacing, and the two-active-dipole antenna pattern is sufficiently similar (as shown in Figure 3.6), then the improvement in

![Figure 3.22](image1.png)

Figure 3.22: Demonstration of antenna pattern diversity with one active dipole and one passive dipole for the a) $\lambda/2$-separated free-space array, b) $3\lambda/4$-separated free-space array, and c) $\lambda_d/2$ array on $\varepsilon_r=16$ hemispherical lens with $\varepsilon_r=4$ matching layer.
channel capacity must come from a combination of better signal decorrelation or
decoupling and antenna pattern diversity. Figure 3.22 shows the improved antenna
pattern diversity offered by the dielectric lens design. Figure 3.22 a) and b) show the
one-active-dipole antenna patterns of the half-wavelength and three-quarters-
wavelength separated two-element free-space arrays, respectively. Considering that if
the alternate dipole is active the patterns would be flipped vertically, it is clear that the
antenna pattern of the lens in Figure 3.22 c) has superior spatial diversity by virtue of
its asymmetry relative to the horizontal axis, while the λ/2 free-space case has the
worst diversity due to large areas of overlap, confirming the results of the channel
capacity graphs.

This verifies the presence of improved antenna pattern diversity in the dielectric lens,
providing one contributing factor to its improved channel capacity. Since the lens
simulates an electrical spacing between elements comparable to λ/2 of free-space
separation, it can be presumed that the decorrelation between the antenna elements is
comparable to the λ/2 free-space array. Finally, the effectiveness of antenna decoupling
can be verified by examining the antenna’s scattering parameters. This is carried out as
part of the antenna characterization in Chapter 4.

3.4 Conclusions

This chapter introduced the proposed lens design and the process used to evaluate and
arrive at the final design from the precursory stages. After establishing a basic antenna
design and process for evaluation of antenna performance, it was possible to improve
the antenna and track the effects of the different design choices through the use of the
parametric study and iterative testing. Adjusting an assortment of physical properties
such as lens size, shape, and material composition in varying combinations allowed for
insight into why the lens behaves as it does, and allowed for educated refinement of the
lens features that lead to the final design put forward. In addition to the design process used to improve the lens’ channel capacity in the ring of scatterers model, the effects of properties such as lens radius, dielectric losses, and lens directivity were investigated and scrutinized.

The iterative process used to test and improve the antenna design resulted in a design that was a hemispherical lens structure consisting of two concentric layers. The first layer was a dielectric material with $\varepsilon_r = 16$ hemisphere and $\lambda_{fs}/4$ radius, and the second layer was a matching cap of $\varepsilon_r = 4$ with $\lambda_{fs}/8$ thickness. The lens was fed with two half-wave dipole strip antennas separated by $\lambda_{lens}/2$. The full lens antenna diameter was $3\lambda_{fs}/4$, and the lens performance was accordingly evaluated against two strip dipoles in free space with separation equivalent to the element separation of the lens ($\lambda/2$) as well as the lens diameter ($3\lambda/4$).

The RoS simulation model provided a valuable method for evaluating the “average” performance of the antenna designs, but the model does not effectively demonstrate the antenna’s behaviour in a more likely “real-world” channel. In the following chapter, the performance of the chosen lens antenna design is further explored through the use of stochastic variation of the RoS model and additional methods of evaluation and analysis.
Chapter 4
Supplementary Tests and Numerical Results

This chapter extends the analysis of the hemispherical lens antenna to support the viability of the idea proposed in this thesis – that a high-permittivity dielectric can be used to reduce spatial correlation between radiating elements with small inter-element separation. The design established in the progression of the parametric study is exposed to a more vigorous battery of tests to determine its efficacy. The original model used in the design stages is modified to provide a stochastic experiment that allows for new metrics to be studied, a full-wave simulation is devised and executed to model a practical communication channel, and additional properties of the lens are reviewed to reinforce the design choices of the previous chapter.

4.1 Variation of the Ring of Scatterers Model

The original ring of scatterers model used in the parametric study is a suitable tool for evaluation of the antenna in the design stages. Since the arrangement of the transmitting and scattering elements remains consistent with respect to the lens and capacity is calculated using receive SNR, the pattern diversity of the antenna becomes the prominent factor affecting channel capacity. Further, with no polarization diversity or signal orthogonality between the transmitting elements, the effect of antenna pattern diversity becomes comparable to spatial decorrelation. Additionally, the uniform scatterer weighting provides a good measure of “average” antenna performance over all angles of incidence, removing any directional bias that would come from the orientation of the transmitting antennas relative to the receiver. This level of consistency and is
useful for direct comparisons between different candidate antenna designs. However, after the optimal design has been selected using the results of the RoS model, it is useful to alter the model to accommodate variation in factors that may affect the wireless channel in order to see how the antenna compares to the benchmarks in different channel environments.

In order to accomplish this, a stochastic version of the RoS model was used. As described in Chapter 2, the original test used a reflection coefficient of -1 to simulate isotropic PEC scatterers. In the new model, both the scatterer reflection coefficient magnitude and phase change of each scatterer in the ring were varied using a uniform random distribution. The magnitude change provided the effect of different multipath signals being reflected by scatterers of different material properties, while the phase change could be interpreted as additional material changes as well as small variations in effective path length travelled. This effect is similar to statistically varying the physical scatterer locations without changing the geometry of the model. These two properties were randomized over several thousand iterations (Monte Carlo simulations) in which the channel capacity for a chosen static receive SNR level was calculated for the lens antenna and the two benchmark free-space dipole arrays. The parameters used in the simulation are outlined in Table 4.1.

Table 4.1: Stochastic RoS Model Test Setup

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$</td>
<td>\Gamma</td>
<td>]</td>
</tr>
<tr>
<td>$\Phi$</td>
<td>Scatterer phase change</td>
<td>[0,2\pi]</td>
</tr>
<tr>
<td>$SNR_0$</td>
<td>Receive SNR measurement point</td>
<td>0 [dB]</td>
</tr>
<tr>
<td>$N$</td>
<td>Number of iterations</td>
<td>10000</td>
</tr>
</tbody>
</table>
Figure 4.1: Histograms showing channel capacity at 0 dB receive SNR over ten thousand Monte Carlo simulations with varying scatterer weighting/phase arrangements and a composite line graph of the lens performance against the two benchmark cases.
The first three charts of Figure 4.1 show the results of the Monte Carlo simulations in the form of histograms that provide counts for channel capacity calculated at the measurement point of 0 dB receive SNR. The test point of 0 dB was used since it falls within the range of receive SNRs used in the original RoS simulation, allowing for a direct comparison between those results and the results of the Monte Carlo simulations (specifically the comparison of median and average capacities at 0 dB SNR). Since the capacity for each iteration of the simulation is calculated using Equation 2.3 and each randomized configuration of scattering elements results in a unique channel matrix $H$, the SNR in the equation acts as a factor that scales the calculated capacity appropriately on a logarithmic scale; there is rapid capacity improvement in the -10 dB to 10 dB range, and moderate improvement over the 10 dB to 20 dB range. The fourth chart in Figure 4.1 provides a composite comparison of the results for the three test cases. It is apparent from the graphs that the lens antenna provides a capacity advantage over the two free-space benchmarks. It can also be observed that the distribution of the $\lambda/2$ free-space array (shown in red) is spread over a wider range of calculated channel capacities while the $3\lambda/4$ free-space benchmark (green) and the lens (blue) have narrower distributions. This may be related to poorer spatial decorrelation at $\lambda/2$ free-space separation, indicating that the arrangement is more sensitive to the changes in scatterer properties. Table 4.2 shows numerical results corresponding to the experiment that generated the charts in Figure 4.1.
Table 4.2: Stochastic RoS Model Test Results

<table>
<thead>
<tr>
<th>Test Case</th>
<th>Capacity (bits/s/Hz)</th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Minimum</td>
<td>Maximum</td>
<td>Average</td>
<td>Median</td>
</tr>
<tr>
<td>λ/2 Free-Space</td>
<td>2.6343</td>
<td>6.3479</td>
<td>4.9433</td>
<td>4.9755</td>
</tr>
<tr>
<td>3λ/4 Free-Space</td>
<td>5.9035</td>
<td>7.6794</td>
<td>6.8395</td>
<td>6.8466</td>
</tr>
<tr>
<td>Hemispherical Lens</td>
<td>6.8731</td>
<td>8.7408</td>
<td>7.9441</td>
<td>7.9528</td>
</tr>
</tbody>
</table>

The adjacency of the average and median values indicates that there were few statistical outliers in the data. Additionally, the mean capacity for each case was lower than the values simulated using the original RoS model in Chapter 3. This can be accounted for by examining the range used to determine each scatterer’s reflection coefficient magnitude. Due to the uniform random distribution used in assigning the values to each scattering element, the average weight of each scatterer was 0.75 (down from 1, the value used in the uniform ring model), resulting in a smaller relative contribution to the total received signal by each multipath signal. The range [0.5,1] was chosen since a magnitude greater than one would imply that the scatterers add energy to the system. Since it is unlikely that many wireless channels would have reflection coefficients as high as the [0.5,1] range, it can be hypothesized that in a channel with even lower reflection coefficient values, the mean channel capacity would be further reduced for all three test scenarios. However, given a sufficient number of reflectors, the lens antenna should still perform better than the two free-space cases. The range chosen for phase variation was [0,2π]. Given these considerations, the results of the Monte Carlo simulations for the two free-space array benchmarks do fall within the
expected range of channel capacities (normally between two and ten bits/s/Hz at 0 dB receive SNR) [26, 27, 31-33].

Since there is now a probabilistic method of evaluating the channel capacity, an additional metric can be added to the assessment of the lens design: the percentage of tests in which each test case provides the highest channel capacity. While Table 4.2 provides results based on the full information provided over the course of ten thousand Monte Carlo simulations, the individual experiments can also be examined independently. For each iteration of the test, the channel capacities of the $\lambda/2$ and $3\lambda/4$ free-space separated dipoles and the $\varepsilon_r = 16$ lens with matching layer were compared. The result was that the hemispherical lens had the best channel capacity of the three test cases in 99.97% of the random scatterer arrangements. From Table 4.2, the maximum capacity of the $\lambda/2$ benchmark was lower than the minimum capacity of the lens, so it can also be deduced that the $\lambda/2$ case was never the best-performing; the implication is that in only three of ten thousand tests the $3\lambda/4$ free-space separated dipoles outperformed the lens.

The results of the stochastic RoS model provide further evidence that the hemispherical lens design is effective in improving channel capacity. Since the normalized antenna patterns of both the lens antenna and the two dipoles with $\lambda/2$ free-space separation are sufficiently similar (as shown earlier in Figure 3.7) when both dipoles are active, this entails that the lens provides significantly improved spatial decorrelation of the two dipoles at close element separation.

4.2 Two-Dimensional Room Simulation

While the Ring of Scatterers model provided an adequate measure of the lens antenna’s average performance in an abstract setting, the results can be further reinforced by
simulating the antenna in a wireless channel that more closely emulates a real-world environment. A two-dimensional “room” was created in SEMCAD X and populated with various reflecting elements in order to simulate a channel with high scatterer density in a full-wave simulation. The lens antenna and two-element free-space arrays were now run in the transmitting mode and signal strength was observed at various points in order to calculate average channel capacity over an area of the room.

4.2.1 Setup

In the development of the full-wave simulation, a two-dimensional environment was chosen over a three-dimensional space for various reasons. Firstly, the RoS model took into account only the H-plane patterns of the antennas being tested, so simulating the

Figure 4.2: Layout of the 2-D full-wave simulation space. PEC and dielectric scattering elements are shown in blue and green, respectively. Scale shown at bottom right.
room in the same plane provides a fair comparison of results. Related to the previous point, since the antenna elements being tested are z-oriented dipoles, the dominant E-field pattern exists mainly in the H-plane, so the inclusion of the third dimension in the simulation would add negligible value to the results and would also require scattering elements to be placed at various vertical depths in the room in order to capture a truly high-density multipath environment, complicating the architecture of the simulation. Finally, the computational time and memory required for a 3-D model would be excessive since SEMCAD X uses the FDTD method, requiring the E-field to be calculated for every voxel over each time step of the simulation. While this simulation technique is effective for the relatively small structure of the lens antenna, it quickly becomes unmanageable in a test environment spanning 100 m$^2$ that would take up approximately 200 m$^3$ if depth was taken into account, even with the aid of hardware acceleration.

A two-dimensional space was simulated in SEMCAD X by limiting the heights of all objects to three cell lengths (375 μm) in the z-direction and exciting the $\text{TE}_z$ mode only. This was accomplished by setting the z-direction boundaries to be PECs and using line sources between the upper and lower limits of the simulation. The x- and y-direction boundaries were set to be strong Absorbing Boundary Conditions (ABC) to limit the simulation to within the room.

The simulation space was constructed using a rectangular room with outer wall measurements of 8 m x 12 m, as seen in Figure 4.2. The outer walls (shown in black) were modeled using dielectric blocks with $\varepsilon_r = 5$, $\sigma = 0.001 \text{ S/m}$, and 0.15 m thickness. The scattering elements were modeled as PEC reflectors (blue) and dielectric blocks (green) with varying physical properties within a predefined set of limits. Table 4.3 shows a summary of the parameters used in setting up the test space.
Table 4.3: Two-Dimensional Simulation Setup

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$l \times w$</td>
<td>Room Dimensions</td>
<td>$8 \text{ [m]} \times 12 \text{ [m]}$</td>
</tr>
<tr>
<td>$l_0 \times w_0$</td>
<td>Sampling Area Dimensions</td>
<td>$4 \text{ [m]} \times 4 \text{ [m]}$</td>
</tr>
<tr>
<td>$\varepsilon_{r,w}$</td>
<td>Wall Relative Permittivity</td>
<td>5</td>
</tr>
<tr>
<td>$\sigma_w$</td>
<td>Wall Conductivity</td>
<td>0.001 [S/m]</td>
</tr>
<tr>
<td>$n_{\text{PEC}}$</td>
<td>Number of PEC scattering elements</td>
<td>6</td>
</tr>
<tr>
<td>$n_d$</td>
<td>Number of dielectric scattering elements</td>
<td>6</td>
</tr>
<tr>
<td>$\varepsilon_{r,d}$</td>
<td>Dielectric Scatterer Relative Permittivity</td>
<td>[1,10]</td>
</tr>
<tr>
<td>$\sigma_d$</td>
<td>Dielectric Scatterer Conductivity</td>
<td>[0.001,0.01] [S/m]</td>
</tr>
</tbody>
</table>

As stated previously, fine gridding and high voxel density results in computationally expensive simulations when dealing with a large space such as the two-dimensional room. However, in order to simulate the chosen antenna design, a high voxel count is required to accurately characterize the curved surface of the lens. In order to circumvent this issue, the lens antenna was replaced with forty point sources, forming a Huygens surface outside of which the E-field is behaviourally identical to the lens antenna. In order to determine the correct calibration of the Huygens sources, the lens antenna was simulated within a two-dimensional box of point sensors (10 cm x 15 cm) with $\lambda/10$ separation between sensors, as shown in Figure 4.3. For the two free-space arrays, two single point sources with appropriate free-space separation were used as the transmitting antenna. The transmitter placement was centered at (0.75 m, 4 m).
A 4 m x 4 m sampling area (as shown in Figure 4.2) was chosen free of any scattering element placement on the opposite side of the room from the transmitter. A field sensor encompassing the area was used to capture E-field strength throughout the sampling region, resulting in approximately 100 000 points of data per simulation, which could later be pared down to create meaningful results in post-processing. The sensor measured total E-field at each data point in the sampling area, providing results comparable to an ideal omnidirectional receiver.

Figure 4.3: Sensor/source setup used to emulate the lens antenna in the two-dimensional simulation.
4.2.2 Evaluation

As with the RoS model, each test case was run twice (alternating the active and passive antenna elements at the transmitter) in order to provide separate impulse responses in the sampling area that could later be used to form the channel matrix and calculate channel capacity of a MIMO system. The field sensor used to measure E-field provided data points with roughly \( \lambda/10 \) separation over the sampling area. In the interest of examining a MIMO system in which the receiver is a two-element array separated by \( \lambda/2 \), every fifth sample was extracted in both the x- and y-directions. This resulted in roughly 4000 individual data points that could be combined with adjacent points to form a dataset representing such a two-element receiver.

In order to more closely match the test conditions to the RoS results, sensor pairings were only considered in the y-direction. This kept the alignment of the receiver matched with the alignment of the transmitter as opposed to if x-directional pairings were considered, in which case the antenna pattern of the receiver pairing would be perpendicular to the transmitter. Recall that in the RoS model a two-element array (in which each element was individually excited by a 1 V source) was used as the transmitter. In order to form the channel matrix, the E-field was measured in SEMCAD X and used in combination with the scatterer geometry and antenna pattern of the various receiving antenna configurations to calculate the total received signal strength based on the summation of all multipath signals at each receiving antenna. In the two-dimensional room simulation, each transmitter arrangement was likewise excited by a 1 V source at one element while the other element was passive. The nature of the full-wave simulation then included all multipath signals in the measurement of E-field over the sampling area. The resulting measurements formed a set of two impulse responses for each transmitter configuration (one for each active/passive element...
combination), and from each of these responses two adjacent test points with appropriate spacing could be selected to provide the measurements required to form the channel matrix $H$ for that particular pairing of test points. That is, the element $h_{mn}$ of $H$ could be extracted by measuring the E-field of the $n$th receiving element in the sampling area for the response generated when the $m$th radiating element of the transmitter is active, and then dividing by the source voltage of antenna $m$:

$$h_{mn} = \frac{E_{\text{meas}}|_{m,n}}{V_{\text{source}}}$$  \hspace{1cm} (4.1)

Since the source voltage used in all experiments was 1 V, the measured E-field at the sampling point was conveniently equal to the corresponding element in the channel matrix, without need for further manipulation.

The channel capacity of each receiver pair was calculated according to Equation 2.3 and the results were averaged over the 4032 unique pairings in order to avoid outlying results such as patches of the sampling area that may have had deeply faded signal strength. The results were then used to generate channel capacity graphs similar to those in Chapter 3.

4.2.3 Results

The channel capacity graph for the two-dimensional room simulation can be seen in Figure 4.4. Notably, the lens antenna with matching layer continued to provide a significant advantage over the two-element free-space arrays used as benchmarks for channel capacity. Figure 4.5 shows histograms of each test setup with counts of capacity range for each of the pairings sampled. It can be noted that the overall distribution of the histograms is similar to that seen with the random RoS test in Figure 4.1, suggesting that the MIMO improvement behaves similarly in both models. It is also apparent that the half-wavelength separation benchmark included a
significant number of “dark spots” visible at the lower end of the histogram distribution. Additionally, it should be noted that the general performance of all three antenna configurations was poorer than the results of the Ring of Scatterers model (dashed lines in Figure 4.4).

This result may be attributed to multiple factors specifically related to the two-dimensional simulation. Firstly, the natural inclusion of path loss in the results of the full-wave simulation calculations would add a scaling element to the E-field contributed by each multipath which was not present in the RoS simulation. Secondly, given the range of relative permittivity values assigned to the dielectric scattering elements in Table 4.3 and the equation to determine reflection coefficient from those multipaths,
Figure 4.5: Histograms showing channel capacity of the benchmarks and lens antenna system over the sampling area of the 2-dimensional full-wave simulation.
\[ \Gamma = \frac{\sqrt{\varepsilon_{r1}} - \sqrt{\varepsilon_{r2}}}{\sqrt{\varepsilon_{r1}} + \sqrt{\varepsilon_{r2}}} \]  \hspace{1cm} (4.2)

the range of reflection coefficients for the multipaths generated by the dielectric scatterers falls within \([-0.519,0]\). Since the magnitude of the reflection coefficients from the dielectric scatterers is noticeably smaller than the ranges used in the modified RoS model, areas of the sensor field that are more dominantly affected by multipath signals from the dielectric scatterers may have lower E-field measurements overall. One example of such a “dark spot” might be the portion of the sampling area in which a PEC reflector partially obstructs the line of sight between the transmitter and sensor field and three prominent dielectric scatterers are located close by. This effect was explored by repeating the stochastic RoS test – in which there are 59 multipath signals and one line of sight signal – with scatterer weights varying in the same range and with fixed phase shift. This revised test showed mean 0 dB channel capacity for all three test cases to be within 6% of the 2D simulation results confirming one of the reasons for the difference in measured results between the RoS model and the room model. Finally, the benchmark test cases saw a significantly more pronounced drop in capacity than the lens antenna between the RoS model and the room simulation. At 0 dB receive SNR, both the \( \lambda/2 \)- and \( 3\lambda/4 \)-separated arrays had a channel capacity roughly 17% lower in the two-dimensional simulation while the lens antenna only performed 12% worse. This disparity could possibly be attributed to the better antenna directivity of the lens antenna. Since the transmitter is positioned and oriented such that there are no scattering elements behind it (with the exception of the walls of the room), the two-dimensional room simulation is not “direction-neutral” like the original RoS model. This can be likened to using a half-ring model in which scattering elements are only placed in the positive-x direction relative to the antenna, resulting in stronger multipath signals to and from the front lobe of the antenna. The result is a channel environment
provides an advantage to the lens antenna due to its increased directivity and penalizes
the two-element free-space arrays due to their pattern symmetry.

Overall, the lens antenna average channel capacity provided a 60% improvement over
the higher benchmark result, demonstrating that the design is viable in a practical
wireless channel with strong multipath components and further substantiating the
claim that a dielectric lens antenna can be used to improve the performance of a MIMO
system that utilizes closely-spaced antennas.

4.3 Further Antenna Characterization

With an extensive collection of simulation data demonstrating the effectiveness of the
hemispherical lens, additional network attributes can be investigated in order to
provide a more complete description of its properties. The antenna characteristics in
this section are the result of post-processing harmonic simulations of the lens antenna
using the design specifications and test parameters outlined in Table 3.1 and Table 3.2,
respectively.

4.3.1 Input Impedance

The source impedance in every instance of strip dipole antennas for the two-element
free-space arrays and for lenses in the parametric study was 50 Ω. This default value
was chosen since it is a common source and load impedance used in RF systems, and
since the input impedance of various lens designs varied. This results in an input
impedance mismatch resulting in a loss in the antenna’s efficiency when examining raw
E-field data, however since in the parametric study channel capacity was calculated
using receive SNR, this mismatch loss would not affect the results of the study.
Analysis of the hemispherical lens with quarter-wave matching layer determined the
input impedance of the final lens design to be 114 Ω. An appropriate matching network
may be added to the antenna feed point in order to tune the input impedance to a more common source impedance, allowing for maximum power transfer.

4.3.2 Antenna Efficiency

As mentioned in the previous section, mismatches between source and input impedance cause a loss in power transferred from the feed source to the antenna, reducing the antenna’s total efficiency. In addition to this mismatch loss, the design of the lens itself and the strength of the dielectric losses in the material chosen also contribute to the antenna’s radiation efficiency, the relation between the power provided to the antenna and the power radiated by or dissipated within the antenna. Equation 4.3 shows the calculation for load reflection coefficient, $\Gamma$, given the measured input impedance $Z_i$ and provided source impedance $Z_0$:

$$\Gamma = \frac{Z_i - Z_0}{Z_i + Z_0} \quad (4.3)$$

The reflection coefficient can then be used to find the mismatch loss, $M_L$, which can be expressed as a ratio or in decibels:

$$M_L = 1 - \Gamma^2 \equiv -10 \log(1 - \Gamma^2) \, \text{dB} \quad (4.4)$$

The antenna’s radiation efficiency, $\varepsilon_R$, is a measured property that can be defined as the ratio of radiated power to input power:

$$\varepsilon_R = \frac{P_{rad}}{P_{in}} \quad (4.5)$$

The mismatch loss and radiation efficiency can then be combined to determine the antenna’s total efficiency:
\[
\varepsilon_T = M_L \varepsilon_R = \left(1 - \left(\frac{Z_l - Z_0}{Z_l + Z_0}\right)^2\right) \left(\frac{P_{\text{rad}}}{P_{\text{in}}}ight)
\]

For the final design of the hemispherical lens antenna with matching layer, the high-dielectric-loss (\(\tan \delta = 0.01\)) radiation efficiency was measured to be approximately 85.13\%. Given the input impedance of 114 \(\Omega\) measured in the previous section and the source impedance of 50 \(\Omega\) from the specification, the total efficiency can be calculated:

\[
\varepsilon_T = \left(1 - \left(\frac{114 - 50}{114 + 50}\right)^2\right)(0.8513) \approx 72.15\%
\]

Table 4.4 shows the radiation efficiency, mismatch loss, and total efficiency for a selection of the \(\lambda_{\text{Lens}}/2\)-separation antenna designs tested during the parametric study.

<table>
<thead>
<tr>
<th>Lens Design</th>
<th>(\tan \delta)</th>
<th>(\varepsilon_R)</th>
<th>(M_L)</th>
<th>(\varepsilon_T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(\varepsilon_r = \mu_r = 8)</td>
<td>0.001</td>
<td>0.411114</td>
<td>0.809768</td>
<td>0.332929</td>
</tr>
<tr>
<td>(\varepsilon_r = \mu_r = 10)</td>
<td>0.001</td>
<td>0.435792</td>
<td>0.890383</td>
<td>0.388026</td>
</tr>
<tr>
<td></td>
<td>0.01</td>
<td>0.280415</td>
<td>0.864480</td>
<td>0.242413</td>
</tr>
<tr>
<td>(\varepsilon_r = 16, \mu_r = 1)</td>
<td>0.001</td>
<td>0.832609</td>
<td>0.838871</td>
<td>0.698451</td>
</tr>
<tr>
<td></td>
<td>0.01</td>
<td>0.710266</td>
<td>0.837400</td>
<td>0.594776</td>
</tr>
<tr>
<td>(\varepsilon_r = 16, \text{Matching Layer})</td>
<td>0.001</td>
<td>0.955712</td>
<td>0.849720</td>
<td>0.812087</td>
</tr>
<tr>
<td></td>
<td>0.01</td>
<td>0.851322</td>
<td>0.847551</td>
<td>0.721539</td>
</tr>
<tr>
<td>(\varepsilon_r = 12, \text{Matching Layer})</td>
<td>0.001</td>
<td>0.977873</td>
<td>0.840107</td>
<td>0.821518</td>
</tr>
</tbody>
</table>
Large aperture antennas (i.e. dish and horn antennas) and half-wavelength dipoles with no lossy materials in close proximity can achieve antenna efficiencies very close to 100%, while antennas used in consumer electronics generally achieve total efficiency in the 35%-70% range due to dissipation caused by surrounding electronics and lossy materials. Since lens antennas are a family of aperture antennas, we expect to see relatively high antenna efficiencies in our design. It should be noted that mismatch loss can be eliminated by properly matching the source and input impedance, so the most important value is the antenna measurements is radiation efficiency.

From the results in Table 4.4, it is apparent that the lenses with matched materials show the worst radiation efficiencies. This can be attributed to the power dissipated within the lens in the form of internal reflections, as discussed in Section 3.3.4. The $\varepsilon_r = 16, \mu_r = 1$ lens had better radiation efficiencies, in the range of 71%-83%. The mediocre efficiency numbers may be attributed to the energy dissipated at the surface of the lens due to the material mismatch at the dielectric-air boundary, also discussed in the same section. Once the matching cap was added, those numbers improved to 85%-95% depending on the strength of the dielectric losses, and even with the inclusion of mismatch loss due to the unmatched input impedance, total antenna efficiency was still above 80%. One point of interest lies in the fact that the $\varepsilon_r = 16$ lens with matching layer was not the design with the highest efficiency: the $\varepsilon_r = 12$ lens with matching layer had a slightly higher radiation and total efficiency measurement. This may be due to the fact that the relative permittivity of the lens and matching layer are

<table>
<thead>
<tr>
<th>$\varepsilon_r = 20$, Matching Layer</th>
<th>0.001</th>
<th>0.936164</th>
<th>0.858537</th>
<th>0.803731</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\varepsilon_r = 25$, Matching Layer</td>
<td>0.001</td>
<td>0.919147</td>
<td>0.856787</td>
<td>0.787513</td>
</tr>
</tbody>
</table>
closer to air, resulting in slightly less power dissipated at the lens surface due to the smaller mismatch.

4.3.3 Gain and Directivity

In the design and evaluation stages, the antenna’s receive SNR was used in the RoS model in order to offset the potential improvement in channel capacity associated with better antenna gain/directivity amplifying the line of sight transmitted signal. Now that a lens design has been selected, the properties of directivity and gain of the chosen design can be reviewed and compared to the other candidate antenna designs.

Directivity is a figure of merit that measures how “directional” the antenna’s radiation pattern is. In general, directivity is a function of angle that is described by the antenna’s radiation pattern. In practice, the term directivity more often refers to the peak directivity, or the power density radiated in the strongest direction compared to the power density radiated by an ideal isotropic radiator radiating the same total power.

General directivity can be expressed using an expression that is a function of angular direction:

$$D(\theta, \phi) = \frac{U(\theta, \phi)}{\left(\int_{\phi=0}^{\phi=2\pi} \int_{\theta=0}^{\theta=\pi} U \sin \theta \, d\theta d\phi\right)/4\pi} \quad (4.8)$$

The numerator is the function of the antenna’s radiation intensity and the denominator represents the antenna’s average radiated power per unit solid angle. Peak directivity, $D$, is the maximum of this function, which corresponds to the spherical coordinate angles in which $U$ is at its maximum. The directivity can then be expressed as a linear value or in dBi (decibels isotropic).
Once directivity and antenna efficiency have been observed or calculated, the antenna gain, $G$, can be calculated. Gain is a measure of power transmitted in the direction of peak radiation compared to an isotropic source after the inclusion of antenna efficiency, and is calculated using,

$$ G = \varepsilon_R D $$  \hspace{1cm} (4.9)

Similar to directivity, gain can be expressed as a function of angle but is more commonly quoted as the gain in the direction of maximum radiation. Radiation efficiency is used instead of total antenna efficiency in the calculation of gain since mismatch loss can be corrected using an appropriate matching network to properly set the antenna’s input impedance. Tables 4.5 and 4.6 show a comparison of peak directivity and gain measurements for a selection of lens designs tested in Chapter 3.

**Table 4.5: Antenna Directivities**

<table>
<thead>
<tr>
<th>Lens Design</th>
<th>D (dBi)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\varepsilon_r = \mu_r = 8$</td>
<td>5.073</td>
</tr>
<tr>
<td>$\varepsilon_r = \mu_r = 10$</td>
<td>5.292</td>
</tr>
<tr>
<td>$\varepsilon_r = 16, \mu_r = 1$</td>
<td>4.778</td>
</tr>
<tr>
<td>$\varepsilon_r = 16, \mu_r = 1, \text{Sphere with Matching Layer}$</td>
<td>3.649</td>
</tr>
<tr>
<td>$\varepsilon_r = 16, \text{Matching Layer}$</td>
<td><strong>6.813</strong></td>
</tr>
<tr>
<td>$\varepsilon_r = 12, \text{Matching Layer}$</td>
<td>6.811</td>
</tr>
<tr>
<td>$\varepsilon_r = 20, \text{Matching Layer}$</td>
<td>4.095</td>
</tr>
</tbody>
</table>
From Table 4.5, it can be confirmed that there was no noticeable correlation between antenna directivity and simulated channel capacity: the directivity of the final antenna design was lower than all of the lenses tested using the cylindrical lens extension of varying lengths, but performed better in the RoS test model. Similarly, the directivity of the spherical lens with matching layer was significantly lower, but the lens still had
higher capacity results than the extended hemispherical lenses. It can also be noted that the $\varepsilon_r = 12$ lens with matching layer had a directivity measurement very close to the $\varepsilon_r = 16$ lens, but the $\varepsilon_r = 16$ lens still showed stronger results. All these examples go on to strengthen the argument that there is little to no correlation between antenna pattern directivity and channel capacity in the channel environment simulated for this study. Similar to the results of Tables 4.3 and 4.4, Table 4.5 shows that the $\varepsilon_r = 16$ lens with matching layer did not have the highest gain of all antennas tested and that the matched material lenses had significantly poorer gains due to their worse radiation efficiencies. These results imply that if transmit SNR was used in the calculation of channel capacity in the RoS design model, the matched lenses may have performed significantly worse and the $\varepsilon_r = 12$ lens with matching layer could have possibly outperformed the $\varepsilon_r = 16$ lens. However, these changes would come as an effect of antenna efficiency and directivity, not a result of improved pattern diversity that we wish to identify in this study.

4.3.4 Scattering Parameters

The scattering parameters, or S-parameters, of an electrical system describe the electrical behaviour of the system and in particular characterize important interactions between the ports of the system. The lens antenna was modelled as a two-port system, in which the ports are the two radiating strip dipole antennas. The scattering matrix for a two-port system,

$$S = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix},$$

(4.10)

shows the relationship between reflected and incident power at each of the dipoles. The antenna design and dipole placement make the system symmetric by observation, resulting in the conclusion that $|S_{11}| = |S_{22}|$ and $|S_{12}| = |S_{21}|$. 
In the case of the lens antenna system, \(|S_{11}|\) represents the antenna self-reflection and \(|S_{12}|\) can be considered a measurement of the mutual coupling between the two dipoles. A multi-simulation setup was employed in which each simulation had one port fed actively while the other port was terminated in a matched load (114 Ω). The results of the simulations were then processed using SEMCAD X’s network analyzer tool to extract the S-parameters. A second set of simulations was run using the two strip dipoles with \(\lambda_{fs}/2\) separation in free-space in which the termination load was 50 Ω. This provided the \(|S_{12}|\) of the \(\lambda/2\) free-space benchmark for comparison to the S-parameters of the lens antenna.

Figure 4.6 shows the S-parameters of the lens antenna and a comparison of mutual coupling between dipoles in the lens antenna and in free-space.

Figure 4.6 shows the S-parameters of the lens antenna as well as the \(|S_{12}|\) of the two dipoles in free-space with half-wavelength separation. While the lens is designed for use at 2.4 GHz, it shows interesting self-reflection and mutual coupling properties around the 2.1 GHz and 2.2 GHz points. However, adjusting the strip dipoles and element spacing for use at those frequencies would result in a completely revised set of S-parameters.
More importantly, when comparing \(|S_{12}|_{\text{lens}}\) and \(|S_{12}|_{\text{fs}}\), it is evident that at the operating point of 2.4 GHz the two values are very close. The lens has a marginally lower \(|S_{12}|\) and should therefore have slightly improved mutual coupling properties. It is normally difficult to separate the impact of spatial decorrelation from reduced mutual coupling. However, if \(|S_{12}|\) is taken to be a characteristic of the antenna that is correlated to mutual coupling between the radiating elements, and since directivity has been shown (earlier in Chapter 3 and the previous sections of this chapter) to have no steady correlation with improvement in channel capacity, then the increase in channel capacity between the two dipoles with \(\lambda/2\) separation in free-space and the lens antenna design must be a result of another factor which can be narrowed down to a combination of improved pattern diversity and spatial correlation: one of the original goals of the design.

4.4 Conclusions

In this chapter, the lens antenna chosen at the conclusion of the design stage was put through more rigorous testing. The original experimental model was modified in order to verify the improved performance of the lens under varying conditions. The antenna was then tested using a full-wave simulation that more accurately modelled a real-world wireless channel. This experiment introduced new factors such as varying relative transmitter-receiver location and the inclusion of path loss into the channel capacity results. In both tests, the hemispherical lens antenna continued to outperform the chosen benchmark two-element arrays, providing reinforcing evidence of the design’s efficacy.

Additionally, the lens antenna was more thoroughly characterized. Properties that are not used as figures of merit were disregarded in the iterative design process, but here they were examined more closely in order to provide further insight into the results of
the channel capacity tests. These results revealed previously hidden information like
the fact that the hemispherical matching layer provided a non-trivial improvement in
radiation efficiency over the basic hemispherical lens, and also provided further
evidence of the previous claim that in this experiment there was no direct correlation
between directivity and channel capacity.

As a whole, this chapter has provided further evidence in support of the previous
design choices and validated the hemispherical lens scheme as a possible solution for
the problem of closely-spaced radiating elements in MIMO systems.
Chapter 5
Conclusions

5.1 Summary

The motivation for this thesis was the exploration of the use of high-permittivity materials as a possible solution to problems associated with close-proximity radiating elements in wireless MIMO systems. While current solutions make use of software to modify signal contents and weighting or hardware additions to improve antenna decoupling and decorrelation characteristics, there are no common antenna designs which attempt to intrinsically reduce signal correlation. The main goals of this study were to design and test the viability of such an antenna, and explore the effect of varying antenna materials and shapes through incremental optimization and varied testing methods.

The well-documented Ring of Scatterers test methodology was combined with an iterative design process investigating the spherical family of lens antennas as a possible candidate for such a solution. Through this process, incremental design decisions were made in order to improve system performance based on the primary metric of channel capacity in a predetermined wireless channel. Physical property variations including lens size, material composition, and shape were considered in this optimization process, and the resulting design was a hemispherical stepped-index lens antenna; an adaptation of an antenna design commonly used in satellite communication applications but not normally considered for localized MIMO systems. Investigating the performance of this design in varied test environments including the two-dimensional full-wave simulation
confirmed that the high-permittivity lens antenna is an effective choice for MIMO systems in rich multipath wireless environments, consistently performing better than the two-element free-space arrays against which it was compared.

In the course of the study, it was determined that a high-permittivity dielectric lens can be used to simulate element spacing greater than the actual physical spacing present between two radiating elements. Further, the design of the lens antenna showed improved channel capacity in a multitude of wireless channel environments. This increase could be attributed to better signal decorrelation due to improved pattern diversity compared to the free-space benchmark test cases. This was an interesting observation since it isolates signal decorrelation as the factor that improved channel capacity; an aspect that is normally tied closely to inter-element coupling. It was possible to arrive at this conclusion due to the sufficiently similar coupling characteristics of the lens antenna and free-space array at the test frequency.

Additionally, there was an interesting observation of the role that antenna directivity plays as a factor in predicting MIMO channel capacity. While much of the optimization process resulted in design changes suggesting that lens antennas with higher directivity had better MIMO channel capacity in the RoS model, testing with various antenna designs preordained to have higher directivities actually resulted in the opposite result, breaking the notion of a direct correlation between antenna directivity and channel capacity.

The results of the design optimization process were confirmed through further testing. A stochastic variation of the Ring of Scatterers model was used to examine the efficacy of the antenna in a pseudo-random environment. The non-uniform wireless channel was then extended to the full-wave simulation of a two-dimensional environment populated with scattering and reflecting elements to ensure a multipath-rich channel. In both
experiments, the results of the tests were consistent with the findings from the design stage: the lens antenna consistently outperformed the benchmark test cases in average channel capacity comparisons.

With a stable of conclusive findings from multiple experimental variations, the chosen antenna design was more completely characterized in order to provide further information and understanding of the previous results. Attributes of the antenna design such as efficiency and directivity – which were not used as metrics in the design study, but can be useful figures of merit – were measured and compared with previous design candidates. These attributes provided additional support for earlier statements and insight into the some of the results from the design process. In particular, additional evidence was found to bolster the argument that directivity does not play as prominent a role in determining MIMO channel capacity as previously thought.

5.2 Future Work

While this study has provided a thorough approach to the design and evaluation of the lens antenna, there is still considerable room for more research and testing. The Ring of Scatterers model used to evaluate early design alternatives forms a very rudimentary and abstract wireless channel. This model is effective when trying to eliminate environmental attributes from influencing the design process, but is not representative of a practical setting. The use of the two-dimensional room simulation brought the validity of the test results one step closer to a real-world scenario, but did not take into account the third dimension and was limited by the tools available to a computational simulation. Thus, the findings of this study could be more powerfully substantiated by fabricating the lens antenna for physical testing in a suitably similar room environment.
Additionally, while the two-element arrays used for benchmark testing throughout the study were chosen to be appropriately comparative to the diameter and element spacing of the lens antenna, they exist in a two-dimensional plane that does not take into account the volume occupied by the lens itself. If physical tests are successful and reveal similar results to the simulations from the study, the next logical step in the design process would be to experiment with lens variations that might allow for a more compact design that would be appropriately matched to the space requirements of MIMO systems that use closely-spaced radiating elements. This step would go a great distance in making such a solution viable in one of the more prevalent areas of MIMO development – mobile wireless applications.

Beyond these two avenues of further development directly related to the lens antenna designed in this study, the design methodology and evaluation processes used here could be carried over to unrelated designs, such as the exploration into the use of metamaterials to construct similarly effective MIMO antennas. The geometry of the RoS model could be easily adapted to support a higher antenna count or multi-dimensional MIMO array. Likewise, the density of measurements in the sampling area of the full-wave simulation would also allow the model to be used for testing more complex systems.
References


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