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Isolation Enhancement for Cylindrical Structure Millimeter-Wave Repeaters

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Isolation Enhancement for Cylindrical Structure
Millimeter-Wave Repeaters

by

B. F. Allen

B.S., University of Colorado Boulder, 2016

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The final copy of this thesis has been examined by the signatories, and we find that both the content and the form meet acceptable presentation standards of scholarly work in the above mentioned discipline.
Allen, B. F. (M.S., Electrical Engineering)

Isolation Enhancement for Cylindrical Structure Millimeter-Wave Repeaters

Thesis directed by Prof. Dejan S. Filipovic

In radio frequency (RF), microwave, and millimeter wave (mmw) repeater applications, the receiver (RX) often needs to detect small power levels while simultaneously transmitting (TX) the same signal amplified significantly without distorting the incoming signal. In some applications, this can be done with time division or frequency division duplexing techniques; however, there are full duplex repeater applications that require simultaneous transmit and receive at the same time and same frequency. These full duplex systems often face the challenge of self-interference caused by the transmitter (TX to RX isolation).

This thesis focuses on enhancing the isolation between two antennas on a repeater system using different impedance surfaces. These surfaces designed to suppress the electromagnetic surface waves that propagate. For this thesis, broadband, mmw repeater systems mounted on cylinders are considered. Both numerical and analytical studies are conducted as well as fabrication and measurements are performed for cylindrical repeater applications.

From a literature survey, we present a general theory on the concept of surface and generalize these impedance surfaces into three types: corrugations, pin bed, and mushroom/patch structures. Uniform depth corrugations are first considered due to their simple geometry and rich history. Theoretical approaches demonstrate how these corrugations work as an impedance surface and discuss the 2:1 frequency bandwidth limitation that is fundamental to their geometry.

To overcome the bandwidth limitation, other analytical techniques are considered where the corrugations are tapered in depth such that they are engineered to choke the surface currents. This technique is shown to work consistently over wide bandwidths and is implemented onto both flat and cylinder ground planes to validate these studies and techniques.

Alternatively, reactive loading of the uniform depth corrugations is shown to lead to band-
width improvement. This eventually directs the studies to the mushroom-type impedance surfaces. These types of surfaces can be easily and accurately fabricated using printed circuit boards which can be designed on conformal substrates making them an ideal candidate for cylinder platform applications. Two types of these mushroom structures are studied in depth in this thesis and both demonstrate high isolation improvement over wide bandwidths in numerical studies and measurement results.
Dedication

To my lovely wife, Shelby. An amazing, beautiful, talented woman whose encouragement, love, and support never wavered to turn me into the best engineer I can be.
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Chapter 1

Introduction and Motivation

1.1 Introduction

In space, aircraft, or any small structure where many antennas are used, the electromagnetic compatibility of multiple systems on the same platform is typically considered an important factor in the design and integration of each of these systems. This thesis considers the most challenging case of a radio frequency (RF) repeater where frequency, time, and polarization multiplexing techniques are not used. In RF repeater systems, a high level of transmitter (TX) to receiver (RX) isolation is required in order to prevent self-interference that causes the signals to become unusable, or to prevent the receiver’s low noise amplifier (LNA) from saturating or even damaging components in the repeater [1–3]. Moreover, in high gain systems, it becomes even more important to achieve high isolation as undesired feedback could result in destructive self-oscillation in the system. These issues hold across the RF spectrum including microwave and millimeter wave (mmw) frequencies which are of particular interest for this thesis. There are several general methods for isolation improvement, some of which are discussed here.

A common method of isolation enhancement when the size of the platform is permissive is by antenna separation [4]. Specifically by moving the transmitter farther away from the receiver, from Friis Transmission Equation [4, 5]:

\[ P_R = P_T G_T G_R \left( \frac{\lambda}{4\pi R} \right)^2, \]

it is expected that the received power, \( P_R \) will decrease by 6 dB per octave (or doubling) of
distance (antenna separation), $R$, where $P_T$ is the transmitter power. $G_R$ and $G_T$ are the linear gains of the receiver and transmitter antennas respectively, and $\lambda$ is the wavelength of operation. In space-constrained platforms, this solution may not be practically achievable if the required antenna separation is greater than the platform allows for. In other words, for the isolation levels to be greater than a specified threshold, (1.1) would require $R \geq R_{\text{min}}$. Problematically, $R_{\text{min}}$ may become so large that it is bigger than the repeater platform in order to achieve the desired isolation.

Another way to enhance the isolation is to redesign the antenna(s) to decrease the line of sight gain between the TX and RX apertures. In other words, the pattern of the transmit antenna can be tailored to have less gain in the direction of the receive antenna and therefore the power at the receiver would be smaller. Due to reciprocity [5, 6], the same holds true for the receiving antenna. As seen in (1.1), the line of sight (LOS) gain product of the antennas has a directly proportional relationship with isolation. Methods of null-steering [7–10], polarization diversity [7, 11–13], or aperture manipulation [3, 11, 14] are some examples of possible antenna redesign. These solutions assume that the system requirements for the antennas are flexible enough to allow for such redesign; however, certain applications may have specific gain, beamwidths, and polarization requirements that will not allow for such flexibility. Alternatively, some systems may allow for frequency- or time-division duplexing; however, these techniques are sub-optimal from the point of spectrum utilization.

A third possible approach to increase isolation is to engineer losses in-between the TX and RX antennas which can be practically realized through the use of an absorber [15]. Both foam absorber [16] and surface absorber (such as MagRAM [17]) can be used. However, absorber can potentially become a problem if the application requires high transmitted power since its loss may create thermal issues. Moreover, environmental conditions of the repeater platform should be considered when using absorber, which could be detrimental to the absorber thus resulting in a decrease in isolation.

This thesis focuses on the isolation enhancement approaches which exploit certain properties
of an impedance boundary condition [6], or more specifically, a high impedance surface [18–20] (HIS) or a reactive impedance surface [21–23] (RIS) between the transmitter and receiver. These

Figure 1.1: Models of common reactive impedance structures including (a) corrugations, (b) pin bed or sometimes called bed of nails, and (c) mushroom structure.

structures present a high impedance to an incident surface wave that can strongly attenuate the propagation of the wave along the structure [18]. These structures are often referred to as reactive impedance surfaces because they exploit designing the surface around having a particular reactive quality (i.e. being inductive or capacitive). These RIS’s are typically categorized into three types: corrugations [19,24–26], pin bed [20,22,27,28] (also called bed of nails), or mushroom structures [19,29,30] and are often referred to as metamaterial surfaces [31–33], metasurfaces [20,34,35], electromagnetic bandgap surfaces [29,36,37], frequency selective surfaces [38–40], artificial conduc-
tors [36, 41, 42], or electromagnetic hard and soft surfaces [31, 43–46] throughout literature. These three RIS categories were modeled for visualization and are shown below in Figure 1.1.

The corrugations in Figure 1.1a are deep ribs or groves in a ground plane. The pin bed (also called bed of nails) in Figure 1.1b is an array of small cylinders. The mushroom structures are shown in Figure 1.1c which are effectively an extension of the pin bed but with added metallic “hats” on the pins. They are typically fabricated using printed circuit board (PCB) techniques which makes them particularly attractive for microwave and millimeter wave applications. In order to create the reactive impedance effect, these structures typically require a periodicity that is much smaller than a wavelength [34], which gives insight into the impedance structure’s size. This thesis investigates how these impedance surfaces can be researched and designed to accommodate broadband mmw repeater systems as well as implemented through practical means.

1.2 Wideband Microwave and Millimeter Wave Repeaters

As discussed earlier, the application (system requirements, power handling, bandwidth, environment, etc.) drives the method used for enhancing the isolation between TX and RX antennas. In this section the application is discussed and the methods of using RIS’s or corrugations are justified. The specific RF repeater considered here is a cylinder with an outer diameter of 6 inches and a space constrained length of 12 inches as depicted in Figure 1.2. The transmit and receive
antennas are mounted on the same side of the cylinder and are 8.5 inches apart measuring from each aperture center. The cylinder is hollow with a quarter-inch wall thickness designed to be directly fit into current cellular base-station towers. Throughout this document, the repeater will be referred to as the Cylinder Repeater System (CRS).

The CRS should be designed to operate from at least 17.5-45 GHz to cover emerging mmw communication standards [47–50] and should be able to support any channel size within this bandwidth. The antennas are dual-linear polarized horns with consistent far-field and impedance behavior from 16-45 GHz described in more detail in [51,52]. Each horn is connected to an orthomode transducer (OMT) and they are oriented in the 45 degree plane (referred to as the diagonal- or D-plane) along the cylinder length. A simplified block diagram of one of the antennas is shown in Figure 1.3 showing these OMT’s are integrated with the coaxial connectors using a waveguide splitter and phase matched lines to excite a turnstile junction. The CRS is expected to have a transmit power of over 100 W and is desired to have at least 80 dB of isolation. The TX to RX

![Figure 1.3: Simplified block diagram of a single antenna for the CRS. Showing that each coax connector is used to excite one of the two linear polarizations. For isolation measurements, one of these ports is terminated and the other is connected to the network analyzer while the same is done for the other antenna.](image-url)
isolation is often composed of three levels: antenna level (passive isolation), analog level (active isolation), and digital level (digital processing) [11]. It is important to note that for this application, the 80 dB isolation specification (spec) would be for the antenna level only. The measured isolation is plotted over frequency in Figure 1.4 along with the measured noise floor.

Noise floor measurements are discussed in Appendix B in more detail, but a large difference between the measured isolation and the noise floor demonstrates confidence in the measurements. The spec is also plotted as a horizontal line at 80 dB to demonstrate the need for isolation enhancement. Furthermore, at least 14 dB of enhancement will be necessary to meet the 80 dB spec throughout the band.

![Figure 1.4: Measured isolation of the CRS and measured noise floor. The 80 dB spec is also plotted as a horizontal line. Due to the used network analyzers upper frequency limitation, the isolation was only measured up to 40 GHz.](image)

While antenna redesign could be considered, the CRS has requirements of at least 15 dBi and a 3 dB beamwidth greater than 25°, which limits modifications that can be done to the antenna itself. Additionally, the cylinder length itself limits the antenna separation; and due to the high output power of the transmitter, an absorber may cause thermal issues associated
with the loss. Moreover, current cellular base-stations are located outdoors, and therefore the unknown environmental impact on such absorbers could be problematic. This is why the method of impedance surfaces is investigated to reduce the coupling on the CRS, and more specifically the design of broadband impedance surfaces in order to consistently improve the isolation over the entire 17.5-45 GHz bandwidth. While the application is very specific, the proposed approaches used in this thesis are general and could be applied to other repeater platforms.

1.3 Explicit Definition of Isolation

To ensure consistency and clarity throughout this thesis, we explicitly define isolation as the ratio of transmitted power, $P_T$, to received power, $P_R$:

$$\text{Isolation} = \frac{P_T}{P_R}$$

(1.2)

This conveniently allows for us to express the isolation in decibels (dB) as the negative of the coupling coefficient ($S_{ij}$ for all $i \neq j$) in dB:

$$\text{Isolation [dB]} = 10 \log_{10} \left( \frac{P_T}{P_R} \right) = -20 \log_{10} |S_{ij}|$$

(1.3)

Consider the two special cases where $P_R = 0$ and $P_R = P_T$. In the former, the isolation has to be infinite for any finite transmitted power and it also will be positive infinite dB, meaning that no energy couples from the transmitter into the receiver. The latter case yields an isolation of 1 or 0 dB which means all the energy couples from the transmitter into the receiver. Equation (1.3) is used throughout this thesis, as the isolation levels measured in dB tend to be much more insightful due to the logarithmic scale. It should also be noted that for (1.3), unless the system has gain, the isolation should always be a positive dB value.

1.4 Background and Uses of Impedance Surfaces in Literature

This section further discusses the use and implementations of impedance surfaces to decouple or isolate two antennas found throughout literature from early works to some of the state of the
art designs. Corrugations have the simplest geometry due to their single dimension periodicity, as evident by Figure 1.1a, and were the first of these impedance surfaces found in literature.

Some of the earliest works found using corrugations specifically for antenna isolation came from the Soviet Union in the early 1960’s [25, 26]; although, corrugations as an impedance surface were introduced as early as the 1940’s [24, 53–55]. Elliott compiled a summary on the theory of corrugations in 1954 [56]. Tereshin came up with analytical surface impedance formulas to decouple two slot antennas in [25] and presented experimental results in [26]. This analytical approach is explained in detail and implemented for the Cylinder Repeater System (CRS) application in Chapter 3. His work was furthered by other Soviet scientists Fel’d and Kyurkchan in the 1970’s [57, 58]. In 1967, Lyon et al. [59] investigated different techniques for reducing the interference of antennas for aerospace applications which included corrugations on the plane between the antennas as well as what they refer to as a “fence” which is the equivalent of a bed of nails. In 1971, Lee and Jones investigated surface wave propagating along two-dimensional corrugated structures and derived a characteristic equation that yield forbidden frequency regions in the dispersion diagram [60]. Takeichi et al. [61] discussed how loading the corrugations can broaden the bandwidth for waveguide applications. In the early 1980’s, King et. al. investigated surface reactances that could be desigend to be independent of incidence angle [22, 27] which led to more investigations into a pin bed or “Fakir’s Bed of Nails” as an impedance surface [28].

In the late 1980’s, Kildal and Lier defined electromagnetic hard and soft surfaces [43, 44, 62] which can be realized as impedance surfaces. These hard and soft surface present characteristics that can either enhance or attenuate the surface waves [44, 45] and could be psychically realized with corrugations [44] or metallic strips [63]. In the 1990’s, interest in what is called “metamaterials” (MM’s) began to grow rapidly, which sparked research into electromagnetic MM’s, frequency selective surfaces (FSS’s), and electromagnetic bandgap (EBG) structures [34]. Uno describes MM’s as [34]: “metamaterials are made of the artificial material that Maxwell equations macroscopic property is really dominant. That is, metamaterials are not ordinary materials made of natural molecules but are periodic structures composed of dielectric, or magnetic, or metallic materials.” A
MM surface (which has also been called a metasurface) can be designed to exhibit desired surface impedances, by manipulating the unit cell of the periodic structure. This is also similar with EBG’s. This concludes that metasurfaces can be classified as a type of surface impedance, or if the losses of the surface are negligible, a reactive impedance surface (RIS). This was eluded to earlier, and throughout this thesis, these types of structures will be called RIS’s as oppose to MM’s. Sievenpiper [18, 19] demonstrated how a high impedance surface can be used to decouple two antennas if placed between the antennas. Since then, there have been many other instances in literature where these impedance structures are used to decouple antennas, some of which may be found in [20,23,35,37,64–70].

With the exception of [20, 23, 68], most of these examples are limited to small bandwidths or even single frequency solutions. The bandwidth of such structures has been analyzed and summarized very well by Kildal and Rajo-Iglesias [30]. Lier et al. have demonstrated techniques to improve the bandwidth through manipulation of the unit cell to achieve a balanced hybrid condition [71–73] and with nonuniform capacitive loading [74] which are considered in this work particularly in Chapter 4. Some other relevant work exploring the decoupling two antennas on cylinder structures can be found in [75–77]. This thesis focuses on extending the bandwidth of corrugations and practical implementations of RIS’s to decouple two antennas over wide bandwidths.

1.5 Thesis Organization and Contributions

The outline of this thesis begins with the concept of surface impedance, some general theory behind impedance surfaces, and utilized tools and modeling techniques in Chapter 2. Analytical, numerical, and experimental approaches to corrugations are presented in Chapter 3, and likewise for the mushroom structures in Chapter 4. Chapter 5 presents a summary of this thesis, extends some tips and lessons learned throughout these studies, and concludes with a discussion on further application and direction.

Two approaches are taken to corrugations, the first being traditional uniform depth corrugations while the second approach extends work done in [25, 26] to taper the depth of the corru-
gations. Analytical expressions fundamental to the geometry of corrugations are derived for both these approaches and numerical results show excellent agreement with the theory and are then further validated through fabrication and measurements. Capacitively loading the corrugations can improve the bandwidth and then be more practically realized as the mushroom structure reactive impedance surfaces (RIS) which due to the wide variety of geometries that these RIS’s can have, derived analytical expressions fundamental to the geometry are not perused; however, they are studied numerically and are validated with fabrication and measurements. A soft surface RIS and an omnidirectional RIS are both investigated and implemented throughout these studies and demonstrated practically for use at millimeter wave frequencies. A performance summary of the impedance surfaces investigated here is shown in Table 1.1; out of the eight results pretested, 5 are measurements and 3 are numerical. Note that there are differences in the antennas used for flat ground plane versus cylinder ground plane as well as the antennas are oriented differently for the respective cases. For the flat ground plane, the antennas are in E-plane, meaning the electric field vector of both TX and RX antennas are parallel; the cylinder ground plane show the results for the antennas in the diagonal plane to better mimic the CRS. This thesis presents different techniques to implement impedance surfaces such as corrugations and RIS’s onto different ground planes. While corrugations and RIS’s have been used for isolation before and the concept is not novel, this thesis looks to connect the theory and practical implementation of such impedance surfaces for use with broadband simultaneous transmit and receive repeater systems in millimeter wave frequencies. This thesis is designed to be used as a handbook for impedance surface design, analysis, fabrication, and measurement and present fair comparisons for different techniques while using the practical application of the CRS as a tangible example.
Table 1.1: Summary of results presented in this thesis for different impedance surfaces on flat and cylinder ground planes. All results are measured unless denoted with an * in which case only numerical results are shown. The “Band” used for determining Maximum, Average, and Minimum Isolation Improvement is the 17.5-45 GHz range of the CRS while the Positive Isolation Improvement Bandwidth considers all measured/simulated frequencies including those that fall below 17.5 GHz ans/or above 45 GHz.

<table>
<thead>
<tr>
<th>Impedance Surface Type</th>
<th>Maximum Isolation Improvement in the Band [dB]</th>
<th>Average Isolation Improvement in the Band [dB]</th>
<th>Minimum Isolation Improvement in the Band [dB]</th>
<th>Positive Isolation Improvement Bandwidth $f_{max}:f_{min}$</th>
<th>Impedance Surface Maximum Thickness [mm]</th>
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</thead>
<tbody>
<tr>
<td>Flat Ground Plane with Antennas Oriented in E-plane</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Traditional Corrugations</td>
<td>27.1</td>
<td>12.2</td>
<td>&lt;0</td>
<td>2.2:1</td>
<td>4.1</td>
</tr>
<tr>
<td>Tapered Depth Corrugations</td>
<td>17.9</td>
<td>13.9</td>
<td>10.3</td>
<td>&gt;2.7:1</td>
<td>4.1</td>
</tr>
<tr>
<td>Connected Patches RIS</td>
<td>36.2</td>
<td>15.4</td>
<td>8.5</td>
<td>&gt;2.4:1</td>
<td>1.57</td>
</tr>
<tr>
<td>Hexagonal Patches RIS*</td>
<td>34.3</td>
<td>17.1</td>
<td>7.7</td>
<td>3:1</td>
<td>1.57</td>
</tr>
<tr>
<td>Cylinder Ground Plane with Antennas Oriented in D-plane</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Traditional Corrugations*</td>
<td>18.1</td>
<td>6.9</td>
<td>&lt;0</td>
<td>2.1:1</td>
<td>4.1</td>
</tr>
<tr>
<td>Tapered Depth Corrugations</td>
<td>29.6</td>
<td>10.9</td>
<td>2.7</td>
<td>&gt;2.6:1</td>
<td>4.1</td>
</tr>
<tr>
<td>Connected Patches RIS*</td>
<td>25.4</td>
<td>9.4</td>
<td>1.7</td>
<td>2.9:1</td>
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<tr>
<td>Hexagonal Patches RIS</td>
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<td>14.1</td>
<td>0.8</td>
<td>&gt;2.6:1</td>
<td>1.57</td>
</tr>
</tbody>
</table>
Chapter 2

Impedance Surface Theory and Modeling Techniques

The use of impedance structures for isolation enhancement is not a new concept; there are multiple examples found throughout literature which were reviewed in the previous chapter. This chapter is included to describe the concept of surface impedance; define special cases such as corrugations, artificial magnetic conductors (AMC), electromagnetic soft and hard surfaces, and electromagnetic bandgap structures (EBG); and discuss numerical modeling approaches to such structures.

2.1 Definition of Surface Impedance

The surface impedance derivation has been described before in several works [18, 19, 78–85] and is most often attributed to Leontovich in 1948 [86]. In fact, the surface impedance concept was used in the 1950’s and 1960’s to measure characteristics of the Earth’s surface in geophysical mapping [84, 85]. In its most simplest form, the surface impedance, $Z_s$, can be described with:

$$\vec{E}_{\text{tan}} = Z_s [\vec{H}_{\text{tan}} \times \hat{n}]$$

(2.1)

where the subscript $\text{tan}$ refers to the tangential (on the surface) components of the electromagnetic fields and $\hat{n}$ is the direction normal to the surface. The surface impedance is a convenient tool for solving Maxwell’s equations when the problem only needs to be solved above a surface in place of solving for the fields both above and beneath the surface. The concept of surface impedance is a general idea and can be applied to many electromagnetic problems. For this thesis, the concept of surface impedance from the perspective of surface waves is primarily considered.
Surface waves are a type of electromagnetic wave that becomes bound to a surface where there is an interface of two dissimilar materials, one of the most common situations is two dielectrics. This surface impedance derivation starts by assuming the waves are bound to some surface with normal \( \hat{n} \), the surface wave is propagating along the longitudinal direction (\( \hat{l} \)), and orthogonal to the transverse direction (\( \hat{t} \)) such that \( \hat{n} = \hat{t} \times \hat{l} \) which is visualized in Figure 2.1. In order for the wave to be bound to the surface, the field should exhibit evanescent behavior along the normal direction, \( \hat{n} \), with decay constant, \( \alpha \), while exhibiting a propagating mode behavior along \( \hat{l} \) with propagation constant, \( \beta \). If considering the TM surface mode, this behavior for the electric field is mathematically described in (2.2) where \( n \) and \( l \) are any coordinates along the normal and longitudinal directions respectively and \( E_0 \) is the initial field strength.

\[
E_l = E_0 e^{-\alpha n - j \beta l} \tag{2.2}
\]

Figure 2.1: Coordinate system used to describe the surface impedance.

Utilizing the Maxwell-Ampere Law, the tangential component of the magnetic field (\( H_t \)) along the surface can be found, and the ratio of \( E_l/H_t \) yields an expression for the surface impedance for the TM mode (2.3). Similarly, if the assumptions above are applied to the magnetic field and
Faraday’s law is applied, an expression for the TE mode surface impedance can be derived (2.4) where \( \epsilon \) and \( \mu \) are the respective absolute permittivity and permeability of the surface at frequency \( f \) in Hertz where \( \omega = 2\pi f \).

\[
Z_s^{TM} = \frac{E_l}{H_l} = j \frac{\alpha}{\omega \epsilon} \quad (2.3)
\]

\[
Z_s^{TE} = -\frac{E_l}{H_l} = -j \frac{\omega \mu}{\alpha} \quad (2.4)
\]

If \( \alpha, \omega, \epsilon, \) and \( \mu \) are assumed to all be positive, real numbers, then it is concluded that the TE surface wave only exists on a capacitive (negative and imaginary) impedance surface. Likewise the TM surface wave only propagates on an inductive (positive and imaginary) impedance surface.

It should be noted that due to the different characteristics of the TE and TM modes, a single surface can present two radically different impedances to each of the modes. In other words, a surface that appears to be capacitive to the TM mode, may also present an inductive impedance to the TE mode. Throughout this thesis, the following convention is used for consistency: the propagation direction (longitudinal component) is defined along the \( y \)-axis while the \( x \)-axis denotes the tangential component, thus making the surface normal along the \( z \)-axis.

Sievenpiper describes HIS’s in detail, including surface wave field descriptions, equivalent circuit models, and a forbidden frequency band which is often referred to as an electromagnetic bandgap (EBG) [18,19]. These works are referenced here to help introduce the concept of HIS’s. A practical realization of a HIS is with shorted metallic, hexagonal patches as shown in Figure 2.2 [19]. If the patches are much smaller than a wavelength, then a wave that is normally incident on the surface will see an impedance consisting of series capacitors (\( C \)) and shunt inductors (\( L \)) which create a parallel resonance at angular frequency \( \omega_0 = 1/\sqrt{LC} \) based on the expression for the impedance of a parallel LC-tank:

\[
Z(\omega) = \frac{j \omega L}{1 - \omega^2 LC} \quad (2.5)
\]

At the pole, \( \omega_0 \) of (2.5), the impedance becomes infinite, which is how the structure achieves “high-impedance” characteristics; additionally, the impedance is purely reactive ((2.5) is purely
imaginary). Below $\omega_0$, the impedance is positive thus an inductive reactance, whereas above $\omega_0$, the impedance becomes negative making it a capacitive reactance. Equation (2.5) is only valid where the approximations of $L$ and $C$ are valid, but as frequency increases, the parasitic capacitance and inductance will change this impedance expression. Moreover at high frequency, the structure will become electrically large and therefore a significant phase change across the unit cell will also make the approximations of $L$ and $C$ invalid.

![Figure 2.2: Sievenpiper’s realization of high impedance surface using hexagonal mushroom patches shorted to the ground plane published in [19].](image)

### 2.2 Special Cases of Surface Impedance

By defining these impedances, four special cases are considered, which can be thought of in a circuit model as combinations of short circuits ($Z = 0$) and open circuits ($Z = \infty$). Table 2.1 summarizes these cases and defines the associated types of impedance surface. The most common of these cases is the Perfect Electric Conductor (PEC), which is usually a good approximation for many metallic surfaces. On a PEC surface, the tangential component of the electric field must be zero from the boundary condition; therefore, $E_t = 0$ must be satisfied on the PEC surface and thus $Z_{sTM} = Z_{sTE} = 0$ from (2.3) and (2.4). Due to this boundary condition, the TE surface
mode cannot exist on a PEC, while the TM surface mode can. This can be additionally proven by taking the limit of the dielectric slab waveguide transcendental equation as the slab thickness approaches zero.

Table 2.1: Special cases of surface impedance defined in terms of $Z_{s}^{TE}$ and $Z_{s}^{TM}$

| Surface Name                  | $|Z_{s}^{TE}|$ | $|Z_{s}^{TM}|$ |
|------------------------------|--------------|--------------|
| Perfect Electric Conductor (PEC) | 0            | 0            |
| Perfect Magnetic Conductor (PMC) | $\infty$   | $\infty$    |
| Electromagnetic Soft Surface | 0            | $\infty$    |
| Electromagnetic Hard Surface  | $\infty$    | 0            |

A Perfect Magnetic Conductor (PMC) is not an intrinsic material found in nature due to the nonexistence of free magnetic charges; however, PMC’s are often convenient in solving electromagnetic problems, and they can be practically realized for a limited bandwidth using periodic structures such as the square mushroom patches shown in Figure 2.3 which resemble the Sievepiper patches in Figure 2.2 [36, 41, 42, 87–90]. These types of structures are often referred to as artificial magnetic conductors (AMC). In general, a PMC will be analogous to the PEC for the magnetic field. A PMC should not allow for the tangential component of the magnetic field to exist on the surface, ($H_{l} = H_{t} = 0$) and therefore $|Z_{s}^{TM}| = |Z_{s}^{TE}| = \infty$ from (2.3) and (2.4). Also analogous to the PEC surface, the TM surface mode does not exist on a PMC, but the TE surface mode can.

The next two types of surfaces are similar to a PMC in the sense that such surfaces are not intrinsic materials, but instead are artificially made to exhibit the desired TE and TM properties. They are referred to as Electromagnetic Soft and Hard surfaces as defined in [43,44]. These names are analogous to acoustics. Simply stated, a soft surface does not allow propagation of either the TE or TM surface modes while a hard surface does allows for such modes to propagate. This can be seen through the boundary conditions and the definitions of the surface impedance in (2.3) and (2.4). Let us first consider a soft surface. To prevent the TE surface mode from propagating, it is desired to present a PEC type of condition to the tangential component of the electric field such
that $E_t = 0$, but to also prevent the TM surface mode from propagating, a PMC condition should force $H_t = 0$. Substitution of these conditions into (2.3) and (2.4) yields $|Z_s^{TM}| = \infty$ and $Z_s^{TE} = 0$.

For a hard surface, it is desired that both TE and TM modes propagate, therefore the TM mode should have a finite, non-zero tangential magnetic field component $H_t$, while $H_l = 0$ due to the definition of the TM mode. Similarly, the TE mode will have a finite, non-zero tangential electric field component $E_t$, while $E_l = 0$ by the definition of the TE mode. Substitution of these conditions into (2.4) and (2.3) result in $Z_s^{TM} = 0$ and $|Z_s^{TE}| = \infty$.

From a practical standpoint, true PMC, soft, and hard surfaces are not achievable over a non-zero bandwidth as is evident by the circuit model (2.5), but they can be approximated under certain conditions [46]. PMC’s can be approximated using AMC’s as mentioned above, while soft and hard surfaces are usually implemented using corrugated structures [44], metallic strips on a dielectric substrate [31, 63, 91], or through strip-loaded mushroom type structures [45]. Some examples have been modeled and are shown in Figure 2.4. Note that a pin bed (or bed of nails), like what is shown in Figure 2.5, can be practically realized using the AMC without the top metallic layer.

As mentioned above, corrugations are a common implementation of soft or hard surfaces

---

Figure 2.3: Example of an Artificial Magnetic Conductor (AMC) practically realized as metal patches.
Figure 2.4: Examples of electromagnetic (a) soft and (b) hard surfaces.

Figure 2.5: Example of a pin bed also called a bed of nails.

depending on the orientation of the propagating wave with respect to the direction of the corrugations [44]. Corrugations engraved along the transverse direction (visualized in Figure 2.6a) will act like a soft surface if the depth of the corrugations are roughly quarter-wavelength. A circuit model is presented in [18], that explains this phenomenon as the input impedance of a shorted parallel plate waveguide that is quarter-wavelength long. A short circuit transformed by a quarter-wavelength will result in an open circuit thus creating an “infinite” impedance effect, satisfying the
soft surface condition for $Z_{S}^{TM}$. Because the corrugations are metal, the TE mode will be shorted satisfying the soft surface condition for $Z_{S}^{TE}$. This technique relies on a quarter-wavelength depth, making corrugations bandwidth limited and are usually considered to have about a 2:1 bandwidth. This is because when the corrugation depth becomes half of a wavelength, the short circuit at the bottom presents a zero input impedance at the top which will cause the corrugated structure to satisfy the conditions of $Z_{S}^{TM}$ and $Z_{S}^{TE}$ for a PEC surface. Techniques are investigated to improve the bandwidth of such corrugated structures in Chapter 3. Similarly, corrugations engraved longitudinally to the propagating wave (shown in Figure 2.6b) will create a hard surface based on the boundary conditions that satisfy $Z_{S}^{TM}$ and $Z_{S}^{TE}$ for hard surfaces. However, in order to trap the surface mode in such a structure, the corrugations should be filled with a dielectric [44]. For the corrugated structures to behave as an impedance surface, the corrugation period should be much smaller than a wavelength in order for the electric and magnetic fields to be uniform on the surface. In [44], Kildal suggests that “More than two corrugations per wavelength are needed.”

Soft or hard surfaces can also be implemented using metallic strips on a dielectric [63] similar to what is shown in Figure 2.4a and 2.4b which can be practically fabricated using printed circuit board (PCB) techniques. These strips can also be electrically connected to a ground plane beneath
the substrate using vias similar to what is shown in Figure 2.3 to manipulate the bandwidth of the surface. This is a result of the additional capacitance (result of the electric field between the strips to the ground plane) and inductance (result of current flowing on the via) added which can shift the resonance frequency $\omega_0$ to lower frequencies. This resonance can be controlled by carefully selecting the periodicity of the strips, thickness of the strips, thickness of the dielectric substrate, and periodicity of the vias if used. Similar to the corrugations, the period of such structures should be much smaller than a wavelength in order for a uniform field to be incident on the surface and create an artificial uniform surface impedance.

AMC's can only exhibit properties of a PMC over limited bandwidths by controlling the shape, size, pattern, periodicity, thickness, or material properties of the unit cell on the right hand side of Figure 2.3. Through changing the physical geometry of the unit cell, almost any TE/TM impedance combination can be realized, which leads to the conclusion that these double periodic structures could result in realizations of soft or hard surfaces as well as AMC's. There are specific TE/TM impedance combinations that result in what Sievenpiper defines as a “forbidden frequency band” [19]. This frequency band is often referred to in literature as an electromagnetic bandgap (EBG) because for frequencies within this bandwidth, the structure does not support either TE or TM modes. This may be similar to a soft surface, but the main difference is that the $Z_{s}^{TM}$ and $Z_{s}^{TE}$ can be finite, non-zero values. By the definitions of $Z_{s}^{TM}$ and $Z_{s}^{TE}$, it is clear that this should only happen in the overlap frequency range where $Z_{s}^{TM}$ is capacitive and $Z_{s}^{TE}$ is inductive. In practice, these periodic structures are not isotropic materials, so the TE and TM surface modes may still appear on the structure if $Z_{s}^{TM}$ is only slightly capacitive and/or if $Z_{s}^{TE}$ is only slightly inductive (i.e. if $Z_{s}^{TM} \approx Z_{s}^{TE} \approx 0$, then the surface no longer will exhibit this EBG properties).

The best visualization of the EBG region of a structure is using the dispersion diagram which plots the frequency over wavenumber.

An example of a dispersion diagram is shown in Figure 2.7. The dispersion diagram demonstrates that there is a frequency region such that neither TE nor TM surface modes exist below the light line; therefore, the surface does not support such modes to propagate and a "bandgap"
Figure 2.7: A dispersion diagram of a mushroom structure similar to one found in [29] demonstrating an electromagnetic bandgap (EBG) where neither TE nor TM surface modes exist below the light line. The frequency is plotted over $k \cdot x$ where $k$ is the wavenumber and $x$ is the unit cell dimension along the $x$-axis.

is defined. The light line on the dispersion diagram represents free space propagation of a wave and has a slope of $c_0 = 1/\sqrt{\varepsilon_0 \mu_0}$, the speed of light in vacuum. If a mode is closely bound to the lower side of the light line (consider $\text{TM}_0$ in Figure 2.7), then it exists in the space directly above the surface [29]. As the TM0 mode bends over (around 30 degrees), the phase velocity of the wave becomes much slower than $c_0$ and the mode becomes strongly bound to the surface. Modes that exist above the light line on the dispersion diagram can radiate away from the surface. These types of modes are often referred to as leaky waves.

### 2.3 Numerical Modeling Techniques for Impedance Surfaces

The two common approaches for numerically analyzing infinite periodic structures: plane-wave scattering (Floquet analysis) and eigenmode analysis [29]. Both approaches are equivalent and can be used interchangeably for the purpose of this thesis as is demonstrated below. Both methods
require the modeling of a unit cell, which can be periodic in N-dimensions: most commonly two ($x$ and $y$), but can easily be more than two, such as the hexagonal mushrooms in Figure 2.2 which is periodic in three dimensions on the same plane. Once the unit cell has been modeled, the sides of the unit cell are set up with periodic boundary conditions (PBC) which force the fields on one side of the unit cell to be equal to that on the opposite side with the appropriate phase shift dependent on the cell size and frequency. In the terminology of the utilized software ANSYS Electronics Desktop (HFSS) [92], this is referred to as Master-Slave boundary conditions.

![Figure 2.8: The simulation setups for (a) the plane-wave scattering and (b) eigenmode analysis. More detail can be found in [29].](image)

For plane-wave scattering, a plane wave is incident using a Floquet Port on the surface and the active scattering parameters are computed which can be transformed into impedance parameters yielding the input impedance. If the input impedance is de-embedded to the actual surface of the structure, it can be interpreted as the surface impedance. The incident wave sees this surface impedance and reflects with some phase shift, which depends on the surface reactance, while the
losses depend on the surface resistance. Usually, these losses are very small and can be ignored in this analysis and the surface can be assumed to be purely reactive. By conservation of energy, the incident wave will reflect off of and transmit through the surface, but if the surface is backed by a ground plane such as the surfaces analyzed throughout this thesis, transmission will not be possible and total reflection can be assumed. In this case the reflection coefficient ($S_{11}$) of the Floquet Port will be the means by which this surface impedance is found. If the unit cell is excited with a TE wave, this analysis will yield $Z_{s}^{TE}$; analogously, $Z_{s}^{TM}$ is computed if the excited wave is TM. Usually, we are only concerned with the fundamental modes (TM$_0$ and TE$_0$), since those will be the primary surface modes encountered throughout this thesis; however, the principle can easily be applied to higher order modes.

In the eigenmode analysis, the unit cell is terminated with a perfectly matched layer (PML). The phase shift across the unit cell is varied and the eigenmode analysis will compute resonances in the unit cell corresponding to the wavenumber and therefore compute the dispersion diagram. By looking into the field vectors of the each of these resonance modes, we can determine if the mode is TE or TM and therefore related to the surface impedance. Care should be taken in ensuring the calculated modes are physical and not simply resonances of the dimensions of the air box and the PML. This can be done by changing these dimensions and comparing the results [29]. These simulation setups are described in detail in [29] and a visualization of both setups are shown in Figure 2.8.

The dispersion diagram and surface impedance can be related using (2.6) and (2.7) to find the respective wavenumbers of each mode:

$$k^{TM} = \frac{\omega}{c_0} \sqrt{1 - \left( \frac{Z_{s}^{TM}}{\eta_0} \right)^2}$$  \hspace{1cm} (2.6)$$

$$k^{TE} = \frac{\omega}{c_0} \sqrt{1 - \left( \frac{\eta_0}{Z_{s}^{TE}} \right)^2}$$  \hspace{1cm} (2.7)$$

where $\eta_0 = \sqrt{\frac{\mu_0}{\epsilon_0}} \approx 377$ $\Omega$ is the impedance of free space. It is important to note that these relations only hold true when the mode can exist, which is demonstrated in the following example.
Figure 2.9: (a) The assumed surface reactance of both the TE and TM modes as given by (2.5) where the blue diamond represents an inductive surface (TM modes exists) and the red circle represents a capacitive surface (TE modes exists), (b) the corresponding calculated dispersion diagram using (2.6) and (2.7), and (c) the physical dispersion diagram that takes into account that TM modes only exist on an inductive surfaces and TE modes only exist on an capacitive surfaces.

Consider a surface that has exactly the same TE and TM surface impedance and follows the impedance profile of an LC tank circuit such that $Z_s = Z_s^{TE} = Z_s^{TM} = (2.5)$ as shown in Figure 2.9a. Let $f_0$ be the resonant frequency and $k_0 = 2\pi f_0/c_0$ is the free space wavenumber of the resonant frequency. If this impedance is substituted into (2.6) and (2.7), the respective
wavenumbers are computed and shown in Figure 2.9b, but special care must be taken to ensure the wavenumbers are physical. Because TE modes do not exist on an inductive surface, for $Z_s > 0$ (denoted with a blue diamond in Figure 2.9a), $k^{TE}$ becomes undefined. Similarly if $Z_s < 0$ (denoted with a red dot in Figure 2.9a), then $k^{TM}$ becomes undefined. The physical dispersion diagram that takes these conditions into account is shown in Figure 2.9c.

These techniques are used throughout the studies conducted in this thesis since they and give good insight into the design of such structures. However, since these approaches assume the structures are infinite in size, they do not exactly model a finite structure placed between two antennas. This means antenna isolation analysis using these techniques, must be performed when the platform, both antennas, and the entire finite structure is modeled together. This can result in computationally large problems due to the electrical size and fine details of these full models. Other approaches that simplify these complex structures are used and discussed throughout this thesis, including Impedance Boundary Conditions (IBC) and hybrid solvers such as the Finite Element/Boundary Integral (FEBI) [93] boundary condition.

### 2.4 Electromagnetic Waves on Curved Surfaces

There has been significant amount of work in electromagnetics to understand the propagation of surface waves along metallic cylinders from the early 1950’s to more recent developments [94–100]. In fact, these studies could be considered as early as Sommerfield with the propagation of electromagnetic surface waves on metallic wires [101,102]; however, we are concerned with cylinders that are considered electrically large, $kR \gg 1$ (where $k = \frac{2\pi}{\lambda}$ is the wavenumber and $R$ is the radius of the cylinder), as opposed to thin wires. For electromagnetic surface waves on cylinders, there are two types of waves that are typically encountered: propagation along the axis of the cylinder and propagation around the axis of the cylinder as depicted in Figure 2.10.

If $kR \gg 1$, the surface waves propagating along the axis of the cylinder are very similar to surface waves propagating on a flat ground plane due to the propagation vector seeing almost no difference between the ground types as only the tangential magnetic field feels the curvature of
Figure 2.10: A depiction of the two types of waves that are typically encountered on cylinders: propagation along the axis of the cylinder (red arrow) and propagation around the axis of the cylinder (blue arrow).

Figure 2.11: (a) Cylinder ground plane with radius $R$ and length $2R$ such that $kR = 28$ and (b) the same cylinder ground plane but “bent” to become flat such that it is $2\pi R \times 2R$ in dimensions. The Hertzian dipole is shown in the center of both ground planes so the fields could be monitored both around and along each ground plane.

the cylinder. This is demonstrated in the example below. For propagation around the cylinder, solving Maxwell’s equation results in two possible solutions: inward travelling waves (towards the center of the cylinder) and outward traveling waves (radiating to infinity) [94, 99]. In general, these surface waves that propagate around the cylinder have solutions that can be represented by Hankel Functions [94, 98] or Airy-Fock Functions [99]. If the radius, $R$, tends to infinity, then the
outgoing waves disappear and the solution of the surface waves matches that of a flat metal ground plane [94, 99], which means the surface wave will locally see a flat metallic ground. For the case of the CRS, $kR = 27.9$ and $71.8$ at the lowest and highest frequencies of operation ($17.5$ GHz and $45$ GHz) respectively. Therefore this approximation is used throughout these studies to simplify the analysis of the impedance surfaces (particularly in the case of the Floquet analysis).

In general, the results presented in this thesis demonstrate that the isolation is very similar for flat and cylinder ground planes confirming this theory. To demonstrate this, a simple numerical
Figure 2.13: Normalized, complex magnitude of the fields propagating on cylinder ($kR = 2.8$ which is no longer much greater than 1) and flat ground planes: (a) Electric field along the ground plane, (b) Magnetic field along the ground plane, (c) Electric field around the ground plane, and (d) Magnetic field around the ground plane. Note that because $kR$ is smaller, but the geometries remained constant, the $x$-axis are scaled differently from the results where $kR = 28$ (Figure 2.12).

test was made where a PEC cylinder ground plane with $kR = 28$ was modeled with an electric Hertzian dipole source to excite the TM surface mode on the cylinder as shown in Figure 2.11. A flat ground plane obtained as an unwrapped $kR = 28$ cylinder was also tested and is also shown. All other parameters of both models were kept the same such as convergence criteria, radiation boundary, probe location, etc.

Just above both ground plane surfaces ($\lambda/50$), there are two probe lines that monitor the
complex magnitude of the total electric and magnetic fields for along and around the ground plane as previously denoted in Figure 2.10. These computed fields are shown in Figure 2.12 demonstrating almost no difference for the fields propagating along a cylinder and only minor differences for fields propagating around the cylinder as compared to the flat ground plane.

Changing the ground planes such that $kR = 2.8$ which is no longer much greater than 1 (Figure 2.13), the fields begin to show significant difference between the ground types. This is a result of the curvature now changing the field vectors significantly enough that assuming the wave propagates similarly is no longer valid. In this case, the Hankel function solutions found in [94,98] should be used. Note that in Figure 2.13 because $kR$ is smaller, but the geometries remained constant, the $x$-axis will be scaled differently from the results where $kR = 28$ (Figure 2.12).

Figure 2.14: The isolation plotted over $kR$ for the cylinder and flat ground planes demonstrating that for $kR \gg 1$, the surface modes are propagating very similarly for the different ground planes.

A final numerical experiment similar to the previous one was performed, but this time the isolation between two antennas were compared. The two quad ridge horns discussed later in this thesis were used to show the isolation between two antennas along the axis of the cylinder (as is the case with the CRS). The models and isolation is presented in Figure 2.14 over $kR$. The plot shows good correlation in isolation across the wide bandwidth demonstrating that for $kR \gg 1$, flat ground
plane approximations are adequately accurate for electromagnetic surface waves. Furthermore, the best agreement is for higher values of $kR$ while for lower values of $kR$, the isolation curves demonstrate more disagreement. This validates the theory that as $kR$ tends to infinity, the cylinder surface wave solutions converge to a flat ground wave solutions.

2.5 Summary

This chapter has set the theoretical framework for this thesis. An introduction to the surface impedance theory and the modeling approaches that are utilized for such structures are discussed. Surface impedance is defined as the ratio of the electric field to the magnetic field on that surface. With respect to the direction of propagation, TE and TM modes are clearly distinguished form the obtained expressions for surface impedance; both TE and TM surface modes can see different impedances on the same physical surface based on this principle. From these two modes, four special cases of impedance surfaces were analyzed: PEC, PMC, Soft Surface, and a Hard Surface. All these can be practically realized with the examples of metal sheets, mushroom patches, transverse metallic strips/corrugations, and longitudinal strips/corrugations, respectively. The dispersion diagram is shown to be a way to visualize how such structures perform with respect to different surface impedances including waves that are bound to the surface versus waves that are free to radiate off of the surface. Additionally, the dispersion diagram allows for an EBG to be defined over a certain “forbidden” frequency range where surface wave modes cannot exist. Numerical techniques that use periodic boundary conditions can solve for different resonances of these periodic structures to generate these plots. Additionally, the periodic boundary conditions will allow for the calculation of the surface impedance based on the reflection coefficient (reflection phase) and the relationship between the wavenumber and surface impedance of such surfaces. In the next chapter, this theory is expanded upon using corrugations as an example where a closed form expression can be derived for the surface impedance, while in Chapter 4, this theory is built upon qualitatively to design more complex reactive impedance surfaces.
3.1 Corrugations as an Impedance Surface

In Chapter 2, corrugations were introduced as a type of impedance surface, that can even exhibit soft-surface behavior over a limited bandwidth. This chapter investigates the analytical methods to analyze corrugations and their implementation as an impedance surface. Fundamental bandwidth limitations and solutions to those limitations are discussed. Two surface impedance equations are derived from different approaches and they are implemented numerically for both flat and cylinder ground planes which are validated using measurements.

3.1.1 Theoretical Description of Corrugations

Corrugations as an Array of Parallel Plate Waveguides

Consider the transverse corrugations in Figure 3.1, which are assumed to be infinite in the $x$-$y$ plane. If the depth of the corrugations goes to infinity, then the geometry would appear to be an infinite array of parallel plate waveguides (PPWG’s) of semi-infinite ($z \leq 0$) length. If the depth is made to be finite and we consider the input of each PPWG to be at the top, then it would make sense to consider each of these PPWG’s to behave as short-circuited stubs in a circuit model. For a TEM-line, short-circuited stub, the input impedance is given by:

$$Z_{in} = jZ_0 \tan(kd) \quad (3.1)$$

where $Z_0$ is the characteristic impedance of the transmission line that makes the stub, $k = \frac{2\pi}{X}$ is
the phase constant of the line, and \( d \) is the depth of the corrugation. In (3.1), if \( kd = n\pi/2 \) for \( n = 1, 3, 5, \ldots \), the input impedance becomes infinite which creates the desired high-impedance effect. Similarly, for \( n = 0, 2, 4, \ldots \), (3.1) becomes zero making the structure appear identical to that of a solid piece of metal with no corrugations. The condition when \( n = 1 \) is of particular interest as \( kd = \frac{\pi}{2} = \frac{2\pi \lambda}{4\lambda} = \frac{\lambda}{4} \) which result in quarter-wavelength corrugations. If metallic losses needed to be considered, then

\[
Z_{in} = Z_0 \tanh(\tilde{k}d)
\]

(3.2)

can be used in place of (3.1) where \( \tilde{k} = \alpha + jk \) is the complex wavenumber with attenuation constant, \( \alpha \). For good conductors, \( \alpha \) can be computed using the skin depth of the metal. This simplified circuit model demonstrates how corrugations can be designed to create a high impedance surface which can enhance the isolation between two antennas.

**Corrugations using the Reflection Phase Approach**

The phase of the reflection coefficient can also be used to show how the corrugations create a high impedance surface. If a wave is normally incident onto the structure in Figure 3.1, then by conservation of energy, the wave must reflect or be dissipated in the metal. For simplicity, we consider the corrugations to be made of a perfect conductor and therefore all the energy must be reflected. If the corrugation depth goes to zero, the structure becomes an infinite PEC sheet and
will reflect the incident wave with a 180° phase shift. As the depth of the corrugations get larger, there will be an associated phase shift, $\Delta \phi_d$, with the depth, $d$. Therefore as the wave is incident on the surface, the reflected wave will have a total phase shift of $\phi = 2\Delta \phi_d + 180^\circ$. The phase shift $\Delta \phi_d$ is multiplied by two because the phase shift happens once on the way down the corrugation and a second time after it has reflected at the bottom and travels back up the corrugation. Therefore, if $\Delta \phi_d = 90^\circ$, then the total phase shift becomes 360° or equivalently 0°, which is the expected reflected wave of a high impedance surface [18]. This model assumes that the majority of the reflections come from the bottom of the corrugations and therefore the wall thickness should be much smaller than the period. For normal incidence, the reflection coefficient and input impedance are related through:

$$\rho = \frac{Z_{in} - Z_0}{Z_{in} + Z_0}$$

(3.3)

which has been visually mapped in the Smith chart provided in Figure 3.2, which simultaneously plots the impedance (as circular contours of constant resistance and reactance) and the reflection coefficient, $\rho$, as a polar plot.

If the corrugations are lossless, then the reflection coefficient will have a magnitude of 1 and will be of the form $\rho = e^{j\phi}$, where $\phi$ is given above as the total phase shift of the reflected wave. The impedance (3.1) must be purely reactive and can be represented as $Z_{in}/Z_0 = jx_{in}$ which is plotted on the $|\rho| = 1$ circle on the Smith chart. Substitution of (3.1) into (3.3) will show that the input reactance $x_{in}$ will rotate clockwise around the Smith chart as denoted in Figure 3.2; however, the reflection phase remains a polar plot and therefore rotates counterclockwise from the purely real axis. Computing the phase of (3.3) will reveal this. If $\phi = 180^\circ$, then $x_{in} = 0$ corresponding to a short circuit and therefore an infinite PEC sheet, whereas $\phi = 360^\circ = 0^\circ$ results in $x_{in} = \infty$ corresponds to a high impedance surface.

This circuit model approximation is only valid if the period of the corrugations is much smaller than the wavelength. If this period becomes electrically large, then the incident wave will have a significant phase shift from one wall to the next, making the reflected wave have non-uniform
phase and therefore invalidating this approximation. Moreover, the impact of the sharp edges of the corrugation walls will result in diffraction of the incident wave (Bragg scattering) such that when the periodicity of the corrugations are electrically large, they will not perform as a uniform surface impedance.

Corrugations using the Concept of Surface Impedance

The above discussion only applies to a normal incidence wave on the corrugations, but this specific type of propagation is not expected when considering the TM surface mode between two antennas located in the same plane of the corrugations. We can then consider a TM wave incident at some angle $\theta$ on the corrugations such that the magnetic field is parallel with the corrugations as shown in Figure 3.3 (i.e. $\vec{H} = H_x \hat{x}$). We again assume the wall thickness is much smaller than the period and that the period is much smaller than the wavelength. The vector $\vec{k}_0$ is used to denote the direction of propagation of the wave. Again, it is expected the wave will reflect with some phase shift associated with the depth of the corrugations.

The calculation of this phase shift is not necessarily trivial, but by approximating the surface as a homogeneous material with effective characteristics $\epsilon_2$ and $\mu_2$, which can be complex, the
reflection coefficient can be found. Using $\epsilon_2$ and $\mu_2$, the effective wave impedance and wavenumber of the surface can also be written as $\eta_2 = \sqrt{\frac{\mu_2}{\epsilon_2}}$ and $k_2 = \omega \sqrt{\epsilon_2 \mu_2}$ respectively. The incident wave can be characterized with wave impedance $\eta_0 = \sqrt{\frac{\mu_0}{\epsilon_0}}$, wavenumber $k_0 = \omega \sqrt{\epsilon_0 \mu_0}$, and phase constant $\beta = k_0 \sin(\theta)$. Since the corrugations will totally reflect the wave, the reflection coefficient can be modelled using Fresnel’s Equation for the reflection coefficient for the case of total internal reflection [6]:

$$\rho = \frac{j \left( \frac{\alpha}{\sqrt{\epsilon_2}} \right) \eta_2 - \eta_0 \cos(\theta)}{j \left( \frac{\alpha}{\sqrt{\epsilon_2}} \right) \eta_2 + \eta_0 \cos(\theta)}$$

where $\alpha = \sqrt{\beta^2 - k_2^2}$. If the definitions of $k_2$ and $\eta_2$ are substituted into (3.4), we find that the first term of numerator and denominator resembles the definition of TM surface impedance (2.3):

$$\rho = \frac{j \left( \frac{\alpha}{\sqrt{\epsilon_2}} \right) - \eta_0 \cos(\theta)}{j \left( \frac{\alpha}{\sqrt{\epsilon_2}} \right) + \eta_0 \cos(\theta)}.$$  

or explicitly using (2.3):

$$\rho = \frac{Z_{sTM} - \eta_0 \cos(\theta)}{Z_{sTM} + \eta_0 \cos(\theta)}.$$  

The reflection coefficient will have a magnitude of unity due to the assumption of total internal reflection and will take the form of $\rho = e^{-j\phi}$. This means that $\phi$ can be found by finding the phase of (3.6):

$$\phi = \angle \rho = -\arctan \left( \frac{2 X_{sTM}^T \eta_0 \cos(\theta)}{(X_{sTM}^T)^2 - \eta_0^2 \cos^2(\theta)} \right)$$
where the TM reactance \( X_{TM}^{s} \) is defined such that \( Z_{TM}^{s} = jX_{TM}^{s} \). Taking the tangent of both sides of (3.7) allows for \( X_{TM}^{s} \) to be solved in terms of the \( \eta_0 \) and the reflection coefficient phase \( \phi \):

\[
X_{TM}^{s} = \eta_0 \cos(\theta) \frac{1 - \sqrt{\tan^2(-\phi) + 1}}{\tan(-\phi)} = \eta_0 \cos(\theta) \tan \left( \frac{\phi}{2} \right).
\] (3.8)

This demonstrates that the TM reactance is related to the reflection phase with a tangent function which is expected from the circuit model, (3.1). In [78], Weinstein demonstrates that through application of the boundary conditions at \( z = -d \), the corrugations can be described using the TM\(_0\) mode characteristic equation:

\[
\frac{\phi}{4\pi} = \frac{d}{\lambda} \iff \frac{\phi}{2} = 2\pi \frac{d}{\lambda} = k_0 d.
\] (3.9)

In [78], a more general formula is presented that accounts for higher order modes, but (3.9) completely describes the corrugations for the fundamental TM mode and shows the reflection phase is only dependent on the geometry of the corrugations. Substitution of the characteristic equation into (3.8) yields:

\[
Z_{TM}^{s} = jX_{TM}^{s} = j\eta_0 \cos(\theta) \tan (k_0 d)
\] (3.10)

which resembles (3.1). It turns out that the substitution of (3.10) into (3.6) shows that the \( \cos(\theta) \) dependence can be factored out and canceled in the reflection coefficient for any \( \theta \). Demonstrating that this \( \cos(\theta) \) dependence is only a result of the incident wave and for any \( \theta \), the wave will see the same reflection coefficient and therefore the same surface impedance with respect to the incident wave’s impedance. Thus corrugation’s surface impedance should be independent of incident angle. This is demonstrated numerically later in this chapter. Equation (3.8) can be generally applied to other surfaces such as the pin bed or mushroom structures; however, the characteristic equation (or reflection phase) of these other structures may exhibit dependence on \( \theta \). It is important to note that (3.10) is the representation of (2.3) for the specific geometry of the corrugations in Figure 3.3 where the TM surface mode is excited from a plane wave incident at angle \( \theta \).

Now, we can reconsider the same problem for a TE wave incident on the corrugations. This can be visualized by switching the \( \vec{E} \) with \( \vec{H} \) and \( \vec{H} \) with \( -\vec{E} \) (duality theorem [6]) in Figure 3.3.
The electric field is now completely parallel with the corrugations independent of \( \theta \). When the wave reaches the corrugations, the tangential electric field must be zero to satisfy the boundary conditions for a PEC. We can then use the Fresnel equation for perpendicular polarization on a conductor to evaluate how the wave will reflect:

\[
\rho = \frac{\eta_2 \cos(\theta) - \eta_0}{\eta_2 \cos(\theta) + \eta_0}
\]  

(3.11)

where \( \eta_2 = \frac{(1+j)\sigma}{\delta_s} \) is the impedance of the conductor. The skin depth \( \delta_s = \frac{1}{\sqrt{\pi f \mu \sigma}} \) describes how deep the wave can penetrate into the conductor with permeability \( \mu \) and conductivity \( \sigma \) at frequency \( f \). By definition, a PEC has \( \sigma = \infty \) therefore \( \eta_2 = 0 \) forcing (3.11) to go to -1. For most good conductors, the conductivity is large enough to approximate the conductor as a PEC. Additionally, the skin depth tells us that a metallic surface is inductive by nature and therefore, the TE surface mode will not propagate along the conductor. This demonstrates that for a TE wave, we can say that \( Z_{sTE} = \eta_2 = 0 \) and that corrugations should act no different to a TE wave than a PEC sheet.

Now we have defined the TE and TM surface impedances for transverse-running, PEC corrugations:

\[
Z_{sTE} = 0
\]

\[
Z_{sTM} = \eta_0 \tan(k_0d)
\]

(3.12)

because of the periodic behavior of the tangent function, at \( d = n\lambda/4 \) for \( n = 1, 3, 5, ... \), the corrugations have electromagnetic soft surface behavior. Additionally, for \( n = 0, 2, 4, ... \), the corrugations act as a PEC sheet.

3.1.2 Some Simple Examples

**Floquet Analysis**

We can now consider the example of a corrugated surface shown in Figure 3.4. The corrugations are defined in terms of a unit cell with parameters: depth \( (d) \), wall thickness \( (w) \), and the gap between the corrugation walls \( (g) \) such that the period of the corrugations can be defined as
$p = g + w$. The Floquet analysis \cite{92} requires the mesh on each pair of walls to be the exact same, so the wall thickness is divided by two and used on each side of the unit cell. In this example,

![Image of corrugated unit cell](image)

Figure 3.4: Unit cell of corrugations considered for the Floquet analysis to determine the TM reactance of the surface.

$g = 0.126\lambda$, $w = 0.002\lambda$, and $d$ is varied from 0 to 0.8\lambda while the fundamental TM mode plane wave is incident on the corrugations. The Floquet analysis setup will compute the reflection coefficient and thus the surface impedance which can be directly compared with (3.1). The first result shown in Figure 3.5 showcases a normally incident TM wave. As seen, excellent agreement between the full-wave Floquet analysis and the analytical result gives the confidence in the FEM analysis to be used for similar problems.

It appears from (3.1) that the TM reactance is independent of incident angle $\theta$, which is tested using the Floquet analysis. The incident angle is tested for 45°, 80°, and 85° and the same analysis is performed. Obtained results for the TM reactance are plotted in Figure 3.6. This was also discussed at the end of the previous section where the TM surface impedance equation has a $\cos(\theta)$ term that is associated with the reflection coefficient, but this $\cos(\theta)$ term cancels with the wave impedance of the incident wave.

Figure 3.6 shows that the magnitude of the TM reactance is in fact independent of incidence angle as expected from (3.1). The numerical results do show a slight shift as $\theta$ changes, which is
Figure 3.5: The TM reactance of corrugations vs corrugation depth \((d)\) for normal incident waves. The analytical expression \((3.10)\) is compared with the Floquet numerical results.

attributed to a few things. The first two possible causes are numerical. In the deembedding of the port, there is possibility of error created by the incidence angle. The quality of the mesh could also create error in the results for each incident angle. The small ripple seen in Figure 3.6c is most likely caused by the mesh. The final possible cause of this shift could be a result of the finite wall-

Figure 3.6: The TM surface reactance of corrugations vs corrugation depth \((d)\) for oblique incident waves plane with incident angles \(\theta = 45^\circ\) (a), \(80^\circ\) (b), \(85^\circ\) (c). The analytical expression \((3.10)\) is compared with the Floquet numerical results.
thickness. There will be some reflections and scattering caused by those edges. For higher incident angles, these edges may be more impactful on the computed reflection coefficient. However, this shift appears to be very insignificant even for large incident angles demonstrating the accuracy of the tool in comparison to the analytical expression for corrugations.

When looking more closely into Figures 3.5 and 3.6, the $x$-axis of the plots could easily be replaced with frequency for a fixed depth, $d$. This is also apparent in (3.1) that if $k_0$ is swept for a fixed $d$, the same response is expected. This is demonstrated using the Floquet analysis where a $0.25\lambda_0$ corrugation depth is fixed at frequency $f_0$. Then the frequency is swept from $0.05f_0$ to $2.5f_0$; however, several variations of wall thickness are tested to demonstrate how a large wall thickness will deteriorate the magnitude of the surface reactance as shown in Figure 3.7. This is because as the wall thickness increases, the corrugations now begin to act as a PEC sheet with small resonant structures cut into it. If considering an incident wave on the larger wall corrugations, the majority of reflections will come from the top of the wall and not the bottom of the corrugation.

Figure 3.7: TM surface reactance for corrugations of various wall thicknesses.
From the surface mode standpoint, corrugations should present a capacitive surface impedance to prevent the TM surface mode from propagating. In Figure 3.5, this occurs when $d/\lambda$ is between 0.25 and 0.5; while in Figure 3.7, this occurs when $f/f_0$ is between 1 and 2, demonstrating the bandwidth limitation of corrugations. Once the corrugations become inductive, they may actually help guide the TM mode and would be detrimental from an isolation standpoint.

**Isolation with a Small Ground Plane**

The degradation in the isolation is demonstrated using the simple model shown in Figure 3.8 of two open-ended rectangular waveguides on a small, finite, PEC ground plane. The dimensions

![HFSS model of the small ground plane with dimensions in terms of $\lambda_0$. The apertures are open-ended rectangular waveguides and the ground itself is PEC. The air box is also shown which the FEBI boundary condition is applied to.](image)
are shown in terms of $\lambda_0$ which corresponds to the design frequency. The waveguides are designed to have a fundamental cutoff at 0.9$f_0$ to ensure the fundamental mode at $f_0$ can propagate through the waveguide and radiate as an aperture antenna would. The radiation box is shown in the figure and the walls of this box were assigned a Finite Element Boundary Integral (FEBI) boundary condition. This model is used as a baseline in order to qualitatively and quantitatively see the impact of the corrugations shown in Figure 3.9. Like the unit cell model, the period of the corrugations is 0.156$\lambda_0$, the wall thickness is 0.03$\lambda_0$, and the depth is held constant at 0.25$\lambda_0$. An additional model that uses an Impedance Boundary Condition (IBC) is also used and is shown in Figure 3.10. The IBC is set up as a rectangle in between the waveguides and the impedance is defined to be purely reactive and is defined with (3.10) with $k_0 = (2\pi f)/c_0$ and $d = \lambda_0/4$ where $f$ is the frequency of the simulation while $\lambda_0$ is held constant. This setup for the impedance boundary is also shown in Figure 3.10.

Figure 3.9: HFSS model of the same ground plane in Figure 3.8, but with corrugations. The side view on the right shows the dimensions of the corrugations in terms of $\lambda_0$.

A qualitative analysis can be performed by looking into the complex magnitude of the electric field as shown in Figure 3.11. The three columns represent the three models (Metal Only, Corrugations, and IBC) and the rows represent the frequency in integer increments of $f_0$. The
fields on both the corrugations and the IBC agree very well and demonstrate that the energy is
directed away from the other antenna as compared with the Metal Only case. This demonstrates
that at $f_0$, the corrugations should improve the isolation. However, for $2f_0$, the energy is no longer
squinted, but appears to move across the corrugations and the IBC almost identically to the Metal
Only case. The same trend is seen for $3f_0$ and $4f_0$ which validates the surface impedance equation
for corrugations follows the form of a tangent function.

While the electric field gives a good visualization and qualitative analysis of the corrugations,
it would be difficult to quantify the fields as they are. To better understand the impact, the port
S-parameters are used to give a clear, quantitative representation of the impact of the corrugations
on the isolation. Moreover, this is convenient because the isolation has been defined in terms of
the port parameters (1.3).

Figure 3.12 shows the isolation is greatly improved at $f_0$ for both the corrugations and the
IBC models as compared to the metal only. The plot also demonstrates the bandwidth limitation
of corrugations to be about 2:1 because around $2f_0$, the isolation is no longer improved but in
Figure 3.11: Cut planes of the models with the complex magnitude of the electric field plotted in dBV/m (scale on right side). The three columns represent the respective model at the top, while the four rows represent frequency in integer increments of $f_0$.

Fact, becomes worse. This is because the corrugations now present an inductive surface impedance as evident from Figure 3.5 which encourages the TM mode to propagate along the surface. The IBC appears to exaggerate this effect, but physically, the IBC is the limit when the corrugation period goes to zero. Moreover, the IBC assumes an isotropic surface impedance in contrast with the anisotropic corrugations. In practice, the corrugations would need to be fabricated with a finite period and finite wall thickness and this limit cannot be met with physical corrugations. To demonstrate this, the period was cut in half and then again up to three times as shown in
Figure 3.12: The isolation on the small ground plane plotted over normalized frequency for the corrugations and the IBC as compared to a metal only.

Figure 3.13 and the isolation for each model is shown in Figure 3.14. These result confirm the hypothesis and it is seen that as the corrugation period gets smaller, the isolation curves approach that of the IBC.

These studies are very insightful into the theory of corrugations, the design of corrugations, and their isolation performance; however, the bandwidth limitations are clear throughout all of these studies. From (3.1) and (3.10), it is also clear that this bandwidth limitation is fundamental to the corrugations and their geometry. To overcome this, there have been several attempts to enhance the bandwidth of such corrugations, specifically discussed in [91]. Another possible way to enhance the isolation between two antennas on the same surface is to design a surface impedance in such a way that the surface impedance forces the surface mode to radiate away from the surface rather than propagate along it [21, 25]. Doing this can be implemented with corrugations very easily because the surface reactance of corrugations can take on any value based on the depth. If this surface can be designed such that the impedance is less dependent on $k_0d$, then the surface can be expected to be more broadband.
Figure 3.13: Side view of corrugations model for different periods. For the period cut in half, there will be double the number of corrugations for the same length, therefore it is labeled as “Corrugations (x2)”.

Figure 3.14: The isolation on the small ground plane plotted over normalized frequency for the corrugations and the IBC for different corrugations periods shown in Figure 3.13
3.1.3 Derivation of Tereshin’s Surface Impedance Equation

In [25], Tereshin, investigated the mutual coupling between two antennas if they were located in a plane of an impedance surface. He practically realized a decoupling impedance surface as corrugations of varied depths in order to attenuate the transverse magnetic field component (i.e. a surface that attenuates the TM propagating surface mode). However, a similar surface could be designed to specifically attenuate the TE modes as well using the same procedure for the tangential component of the electric field. Taking into account that the paper may not be easily accessible, the derivation presented in [25] of the surface impedance is also presented here. Note that this derivation is general and is later practically realized using corrugations.

Consider a TM mode that propagates along the $y$-axis on a surface which lies in the $x$-$y$ plane and has a normal parallel to the $z$-axis, such that the transverse magnetic field component, $H_x$, assuming a time-harmonic ($e^{j\omega t}$ which is not shown throughout this derivation) dependence, can be described as the non-uniform wave:

$$H_x = H_0 e^{f(y,z)} = H_0 e^{f_R(y,z) + j f_I(y,z)} \quad (3.13)$$

where $H_0$ is the initial magnetic field strength at the origin and $f(y, z)$ is some complex function representing the attenuation and propagation of the magnetic field. Both $f_R(y, z)$ and $f_I(y, z)$ are real functions that only depend on $y$ and $z$ and the field is assumed to be independent of the transverse direction $x$. Maxwell-Ampere’s law (3.14) can be used to solve for the longitudinal component of the electric field, $E_y$.

$$E_y = -\frac{j}{\omega \epsilon} \frac{\partial H_x}{\partial z} \quad (3.14)$$

where $\epsilon$ is the absolute permittivity of the medium above the surface. Using the ratio of $E_y$ to $H_x$ yields an impedance which resembles (2.3) and is considered a surface impedance if computed on the surface ($z = 0$):

$$Z(y, z = 0) = \frac{E_y}{H_x} \quad (3.15)$$

If (3.13) and (3.14) are substituted into (3.15), an impedance as a function of distance ($y$) can be
computed which only depends on the function $f(y, z = 0)$.

$$Z(y) = \left( -\frac{j}{\omega \epsilon} \frac{\partial f_R}{\partial z} + \frac{1}{\omega \epsilon} \frac{\partial f_I}{\partial z} \right) \bigg|_{z=0}. \quad (3.16)$$

From (3.16), a condition for a purely reactive impedance surface can be found:

$$\frac{\partial f_I}{\partial z} \bigg|_{z=0} = 0. \quad (3.17)$$

Knowing (3.16) and (3.17), a purely reactive impedance surface that attenuates the magnetic field in a specific way can be designed over a limited distance ($y$ going from TX to RX). One solution is if $f_R(y, z = 0)$ is made to be extremely large and negative, which will result in a large attenuation. To practically realize a large attenuation in the surface impedance, (3.16) shows that a large derivative of the function, $f_I(y, z = 0)$, should be used. Assuming the source (TX antenna) is located at $y \leq 0$ and no other sources are present on the surface, then $H_x$ must satisfy the homogeneous wave equation for all $y > 0$:

$$(\nabla^2 + k^2)H_x = 0 \quad (3.18)$$

where $k = 2\pi/\lambda$ is the wavenumber, note that $\frac{\partial H_x}{\partial y} = 0$ as stated earlier, thus simplifying (3.18) to:

$$\frac{\partial^2 H_x}{\partial y^2} + \frac{\partial^2 H_x}{\partial z^2} + k^2 H_x = 0. \quad (3.19)$$

Substitution of (3.13) and (3.17) into (3.19) and separation of the real and imaginary parts yields the system of equations:

$$\begin{align*}
\frac{\partial^2 f_R(y,z)}{\partial y^2} + \frac{\partial^2 f_R(y,z)}{\partial z^2} + \left( \frac{\partial f_R(y,z)}{\partial y} \right)^2 + \left( \frac{\partial f_R(y,z)}{\partial z} \right)^2 - \left( \frac{\partial f_I(y,z)}{\partial y} \right)^2 - \left( \frac{\partial f_I(y,z)}{\partial z} \right)^2 + k^2 &= 0 \\
\frac{\partial^2 f_I(y,z)}{\partial y^2} + \frac{\partial^2 f_I(y,z)}{\partial z^2} + 2 \left( \frac{\partial f_R(y,z)}{\partial y} \right) \left( \frac{\partial f_I(y,z)}{\partial y} \right) + 2 \left( \frac{\partial f_R(y,z)}{\partial z} \right) \left( \frac{\partial f_I(y,z)}{\partial z} \right) &= 0. \quad (3.20)
\end{align*}$$

When considering the solution to (3.20), it is valid to assume that the propagating field is a surface mode such that the longitudinal component of the field resembles (2.2). In this case, we are only concerned with points that are close to the surface (i.e. $z \approx 0$). This allows for $f_R(y, z)$ and $f_I(y, z)$ to be represented by their respective truncated Taylor series expansions up to the linear terms [25]:
\[ f_R(y, z) \approx f_R^{(0)}(y) + f_R^{(1)}(y)z \]  
\[ f_I(y, z) \approx f_I^{(0)}(y) + f_I^{(1)}(y)z. \]  

In (3.21), \( f_{R,I}^{(i)}(y) \) represents the \( i \)th partial derivative of function \( f_{R,I}(y, z) \) with respect to \( z \).

Through substitution of the respective partial derivatives in (3.21) into (3.20) and using the purely reactive condition (3.17) allows for the system of equations to be written as:

\[ \frac{d^2 f_R^{(0)}(y)}{dy^2} + \left( \frac{df_R^{(0)}(y)}{dy} \right)^2 - \left( \frac{df_I^{(0)}(y)}{dy} \right)^2 + \left( f_R^{(1)}(y) \right)^2 + k^2 = 0 \]  
\[ \frac{d^2 f_I^{(0)}(y)}{dy^2} + 2 \left( \frac{df_R^{(0)}(y)}{dy} \right) \left( \frac{df_I^{(0)}(y)}{dy} \right) = 0. \]  

Then, (3.16) can be written as:

\[ Z(y) = -\frac{j}{\omega c} f_R^{(1)}(y) \]  

and therefore, the system of equations (3.22) becomes:

\[ f_R^{(1)}(y) = \pm \sqrt{\left( \frac{df_I^{(0)}(y)}{dy} \right)^2 - \left( \frac{df_R^{(0)}(y)}{dy} \right)^2 - k^2} \]  
\[ \frac{d^2 f_I^{(0)}(y)}{dy^2} + 2 \left( \frac{df_R^{(0)}(y)}{dy} \right) \left( \frac{df_I^{(0)}(y)}{dy} \right) = 0. \]  

If considering the first equation, the first term under the square root can be solved using the second equation of (3.24):

\[ \frac{d^2 f_I^{(0)}(y)}{dy^2} = -2 \left( \frac{df_R^{(0)}(y)}{dy} \right) \left( \frac{df_I^{(0)}(y)}{dy} \right). \]  

We define \( g(y) = \frac{df_I^{(0)}(y)}{dy} \) and \( h(y) = \frac{df_R^{(0)}(y)}{dy} \), then (3.25) becomes

\[ \frac{dg(y)}{dy} = -2h(y)g(y) \]  

which is a first-order linear ordinary differential equation that has solution:

\[ g(y) = C_1 e^{-2\int h(y)dy} \]  

where \( C_1 \) is an arbitrary constant. This also demonstrates that \( f_R^{(0)}(y) \) and \( f_I^{(0)}(y) \) are not independent functions, but are related through:

\[ \frac{df_I^{(0)}(y)}{dy} = C_1 e^{-2f_R^{(0)}(y)}. \]
Equation (3.28) can simplify the first part of (3.24) and be substituted into (3.23) to the desired surface impedance 3.29 where the constant $C = C_1^2$:

$$Z(y) = \pm \frac{j}{\omega \epsilon} \sqrt{Ce^{-4f_R^{(0)}(y)} - \frac{d^2 f_R^{(0)}(y)}{dy^2} - \left( \frac{df_R^{(0)}(y)}{dy} \right)^2 - k^2}. \quad (3.29)$$

Here, we are allowed to choose the sign of the first equation due to the square root. We consider that a positive sign will result in an inductive surface whereas the negative sign should result in a capacitive surface. Even though a capacitive surface should completely eliminate the TM surface mode, we choose the positive with the consideration of the application. The application of the Cylinder Repeater System (CRS) has only quarter-inch thick aluminum, and a capacitive surface impedance will require each corrugation to be at least quarter-wavelength long (Equation (3.29) would be starting from quarter wavelength and changing from there to remain capacitive) whereas an inductive surface impedance can start at a corrugation depth of zero. For this purpose, the positive sign is kept in (3.29). Tereshin’s Surface Impedance Equation is implemented in the form of corrugations in this thesis; however, (3.29) is a general expression for surface impedance and it could be applied to other types of surfaces such as a pin bed or the mushroom structures presented in Chapter 4.

### 3.1.4 Discussion on Tereshin’s Surface Impedance Equation

From Tereshin’s surface impedance equation (3.29), a relationship between the surface reactance and TM surface mode attenuation can be formulated. We take (3.29) and present an expression for the surface reactance:

$$X_s(y) = \pm \frac{1}{\omega \epsilon} \sqrt{Ce^{-4f_R^{(0)}(y)} - \frac{d^2 f_R^{(0)}(y)}{dy^2} - \left( \frac{df_R^{(0)}(y)}{dy} \right)^2 - k^2}. \quad (3.30)$$

Now, we assume that the magnetic field decays exponentially as:

$$H_x(y) \bigg|_{z=0} = H_0 e^{-\alpha y} \quad (3.31)$$
where \( \alpha \) is the attenuation factor (units of Np/m) which can be related to a designed/desired attenuation of \( A_{\text{dB}}/\lambda \) in units of \( \frac{\text{dB}}{\lambda} \) using the next few equations:

\[
-\alpha y = -\frac{\alpha}{k} ky = -2\pi \frac{\alpha}{k} \frac{y}{\lambda}.
\]  (3.32)

Knowing that the attenuation can be found using \( 20 \log_{10} \left( \frac{H_x(y)}{H_0} \right) \bigg|_{z=0} \) and making use of the relationship: \( \log_{10}(e^{-x}) = -x \ln(10) \), we can find the attenuation factor in dB:

\[
A_{\text{dB}}/\lambda = 20 \log_{10} \left( \frac{H_x(y)}{H_0} \right) \bigg|_{z=0} = 20 \log_{10} (e^{-\alpha y}) = -20 \frac{2\pi \frac{\alpha}{k} \frac{y}{\lambda}}{\ln(10)} \]  (3.33)

and therefore the required factor, \( \frac{\alpha}{k} \), can be found for a desired attenuation/wavelength, \( A_{\text{dB}}/\lambda \):

\[
\frac{\alpha}{k} = A_{\text{dB}}/\lambda \cdot \frac{\ln(10)}{2\pi 20}.
\]  (3.34)

Taking into account that \( \frac{1}{\omega_0} = \frac{\eta_0}{k} \), (3.30) can be simplified in terms of \( \frac{\alpha}{k} \):

\[
X_s(y) = \pm \eta_0 \sqrt{\left( \frac{C_1}{k} \right)^2 e^{4\alpha y} - \left( \frac{\alpha}{k} \right)^2 - 1}.
\]  (3.35)

the constant, \( C_1 \), can be found from a known surface reactance, \( X_0 \), at some point \( y = y_0 \):

\[
X_0 \equiv X_s(y = y_0) = \pm \eta_0 \sqrt{\left( \frac{C_1}{k} \right)^2 e^{4\alpha y_0} - \left( \frac{\alpha}{k} \right)^2 - 1} \]  (3.36)

and therefore solved for \( \frac{C_1}{k} \):

\[
\left( \frac{C_1}{k} \right)^2 = \left[ \left( \frac{X_0}{\eta_0} \right)^2 + \left( \frac{\alpha}{k} \right)^2 + 1 \right] e^{-4\alpha y_0}.
\]  (3.37)

The right hand side of (3.37) is a known constant and therefore, can be re substituted into (3.35):

\[
X_s(y) = \pm \eta_0 \sqrt{\left( \frac{X_0}{\eta_0} \right)^2 e^{4\alpha(y-y_0)} + \left[ \left( \frac{\alpha}{k} \right)^2 + 1 \right] \left( e^{4\alpha(y-y_0)} - 1 \right)},
\]  (3.38)

which gives us an expression for the surface reactance engineered for a specific attenuation/wavelength, \( \frac{\alpha}{k} \). Again, note that the \( \pm \) gives us the freedom to decide how to implement this surface reactance as either inductive (+) or capacitive (−). We plot the attenuation over \( y \) as well as the

\[\text{Note that the following logarithm convention is used: } \log_{10}(x) \text{ is the logarithmic function with base 10 while } \ln(x) \text{ is the natural logarithm with base } e.\]
required surface reactance (magnitude of (3.38)) over $y$ for different values of this attenuation/wavelength parameter in Figure 3.15 with the assumptions of $X_0 = X(y = y_0 = 0) = 0$.

Figure 3.15: Using (3.38), the desired attenuation over wavelength (left) can be related to the required surface reactance (right).

Figure 3.16: Small ground plane model to demonstrate Tereshin’s surface impedance equation (3.29) implemented with impedance boundary conditions. A H-plane horn antenna is used to excite the TM surface mode on the ground plane.
We now attempt to validate this design approach numerically with the model shown in Figure 3.16 which has a small ground plane similar to the test bench model used throughout this thesis, but made wider and instead of a single open ended waveguide, an H-plane horn is used with normalized dimensions specified in the model. The formula (3.38) is implemented using the Impedance Boundary Condition (IBC) in HFSS [92]. To simplify the model, equation (3.38) and Figure 3.15 are discretized using \( \lambda/10 \) steps to create forty separate IBC’s, which is demonstrated with a few of the IBC’s in Figure 3.16.

**Inductive Surface Reactance**

First, an inductive version of (3.38) is analyzed by choosing the positive sign before the square root. This inductive reactance is substituted into the IBC of the model. The surface current density can be plotted on the IBC’s and is shown in Figure 3.17. It is expected is that while the surface reactance is low enough, the TM mode will propagate along the surface, but at some point this surface reactance will become too high that the surface current will be choked due to its high impedance value. This effect is seen very clearly in the plot. The threshold impedance for each of these is stated on the plot to show that the impedance needs to be at least 3000\( \Omega \) in order to choke the surface current out completely.

To quantitatively understand how this surface impedance is working, we examine the transverse magnetic field component \( H_x \) just above the surface (at a height of \( \lambda/50 \)) starting from \( X_0 \) in Figure 3.18. The magnetic field does not decay in the expected way from Figure 3.15, instead the magnetic field appears to propagate until it reaches the threshold impedance and then drops drastically, independent of the attenuation, \( (\frac{\alpha}{k}) \). This threshold appears to be around \( X_s = 10\eta_0 \) which we can define as our high impedance threshold (HIT). We can take the impedance plotted in Figure 3.15 and plot it on the Smith chart normalized to \( \eta_0 \) to see what is going on for the different cases of \( \frac{\alpha}{k} \). The blue dots represent the discritized version that is implemented on the IBC in the model above and the HIT is plotted as a green square. Note that the y-axis moves around the Smith chart clockwise when the surface reactance is inductive. As expected, the surface impedance is getting pushed to an open circuit (infinite impedance), therefore the conclusion here is that what
Figure 3.17: Surface current density for the inductive implementation of (3.38), also showing that at a certain point along the surface, the current is choked out.

\[ X_s[y_k = y_{35}] = 3381 \ \Omega \quad \text{and} \quad X_s[y_k = y_{27}] = 4050 \ \Omega \]

\[ X_s[y_k = y_{16}] = 3656 \ \Omega \quad \text{and} \quad X_s[y_k = y_3] = 4088 \ \Omega \]

Figure 3.18: Decay of the transverse magnetic field component along the inductive surface.
appears to be most important in suppressing the TM surface mode is presenting a high enough impedance at some point along the surface that chokes out the surface current.

![Figure 3.19: The inductive surface reactance represented on a Smith chart for visualization of how the surface chokes the TM surface mode. The green square indicated $X_s = 10\eta_0$ while the blue dots correspond to the discretized version of (3.38)](image)

**Capacitive Surface Reactance**

We perform a similar analysis, but choosing the negative sign of the square root in (3.38) to get a capacitive surface reactance. Again, the surface currents can be plotted on the IBC as is shown in Figure 3.20 and the magnetic field decay is plotted over $y$ in Figure 3.21. What is seen is that the surface current is much more dampened than what was shown in the inductive surface reactance case. This is because the capacitive surface inherently suppresses the TM surface mode which is no further enhanced by the high impedance surface. As a result, it is much harder to define a threshold impedance for this case as there is not necessarily a clear cutoff point. This is also very evident in the magnetic field decay in Figure 3.21. The Smith charts show us a similar story to the inductive surface reactances as the capacitive and inductive surface reactances are the mirror image of each other. The only difference is that the $y$-axis moves counterclockwise around the Smith chart for the capacitive surface.

Therefore, what we can conclude that from the perspective of attenuating the surface wave,
Figure 3.20: Surface current density for the capacitive implementation of (3.38), also showing that at a certain point along the surface, the current is choked out.

Figure 3.21: Decay of the transverse magnetic field component along the inductive surface.
Figure 3.22: The capacitive surface reactance represented on a smith chart for visualization of how the surface chokes the TM surface mode. The green square indicated $X_s = 10\eta_0$ while the blue dots correspond to the discretized version of (3.38).

The impedance should change from low to high at some point on the surface to choke out the surface currents. With the capacitive surface, an additional amount of attenuation is achieved from the nature of not supporting TM propagation. From the standpoint of using this surface to create a broadband impedance surface designed to suppress the surface modes from propagating, the surface should be designed to always have an impedance be either capacitive or inductive beyond the HIT. If the surface has impedances that meet these criteria over a large frequency bandwidth, the surface should effectively attenuate the surface mode over that bandwidth.

### 3.1.5 Implementation of Tereshin’s Surface Impedance Equation

The previous two sections provide an analytical expression to the surface impedance. If the designer knows exactly how they want the magnetic field (TM surface wave) to behave, (3.29) provides the exact surface impedance profile with respect to distance needed to accomplish this behavior. In [26], Tereshin implemented his derived expression as corrugations starting with a simple way to describe the magnetic field from (3.13) which was considered in the previous section:
\[ H_x \bigg|_{z=0} = H_0 e^{f(y,z)} \bigg|_{z=0} = H_0 e^{-\alpha y}. \]  

(3.39)

This shows that he desired for the TM surface mode to be attenuated only because the function \( f(y,z) \) has a real part real implying attenuation with constant \( \alpha \) and zero imaginary part implying no propagation. Substitution of \( f_R(y,z) = -\alpha y \) into (3.29) yields:

\[ Z(y) = \frac{j \omega \epsilon}{\omega \epsilon} \sqrt{C e^{4 \alpha y} - \alpha^2 - k^2}. \]  

(3.40)

Equation (3.40) is a very simple expression which tells us that the impedance will vary similar to the desired magnetic field with an added constant. An interesting case is if \( C e^{4 \alpha y} \gg \alpha^2 + k^2 \), then \( Z(y) \approx \frac{j \sqrt{C e^{2 \alpha y}}}{\omega \epsilon} = \frac{\zeta H_0^2}{H_T} \) where \( \zeta = \frac{j \sqrt{C}}{\omega \epsilon} \) is an imaginary constant at any fixed frequency. Another interesting case is if the arbitrary constant \( C = 0 \), then \( Z(y) = -\frac{\sqrt{\alpha^2 + k^2}}{\omega \epsilon} \) which is a purely real, negative, constant surface impedance, thus requiring external power (an active surface), which is not practical for a repeater platform that may already have power constraints.

Equation (3.40) is not frequency independent as is clear from the \( 1/\omega \) and \( k^2 \) terms; however, if we let \( C = \sigma k^2 \) and \( \alpha = a k \), where \( a \) and \( \sigma \) are frequency independent constants, then (3.40) could be simplified to:

\[ Z(y) = j \eta_0 \sqrt{\frac{\mu R}{\epsilon R}} \sqrt{\sigma e^{4a\pi y/k} - a^2 - 1} \]  

(3.41)

where \( \epsilon_R \) and \( \mu_R \) are the relative permittivity and relative permeability of the surface respectively.

Equation (3.41) now shows the only frequency dependence is in the exponential under the square root. An exponential dependence on frequency does mean the magnitude of the impedance changes more rapidly with frequency, but does not have periodic behavior like the tangent function of (3.10), which potentially allows for more broadband performance. It was observed that if \( \alpha = 0.475k \) and \( C = 1.22k^2 \) (which are found in [26]) for \( k \) being the free space wavenumber of 18 GHz, the magnitude of the impedance was so large, it could only be practically realized with quarter-wavelength corrugations covering the entire range of \( y \) (see Figure 3.23). Tereshin was only interested in designing at a single frequency \( (k) \), and therefore his analytical expression to decouple two antennas derived from Maxwell’s Equations should result in a high impedance plane.
The impedance profile of Figure 3.23 demonstrates that for relatively small distances ($y$ on the order of $1\lambda$), (3.40) simply yields a high impedance surface. The frequency dependence of the surface can be analyzed by substitution of different values of $\lambda$ into (3.41) for the fixed values of $a$ and $\sigma$ from the previous plot and is shown in Figure 3.24. As the distance $y$ grows, all frequencies converge to the same high impedance which makes sense due to the exponential nature of (3.29). The only difference in frequency occurs for very small value of distance ($y \leq 0.25\lambda$) which is also shown in the plot.

Figure 3.24 concludes that for long decoupling structures, the surface needs to present high impedance for all frequencies, which is not practical for quarter wavelength corrugations due to the bandwidth limitations thereof. In order to generate an impedance surface with corrugations for broadband behavior, a modified version of (3.29) was used that was discovered through accidental means. This modification is presented in the next section.

### 3.1.6 Modified Tereshin Surface Impedance Equation

Through error in the beginning of these studies, it was observed that by substituting $k' = 1/\lambda$ instead of $k = 2\pi/\lambda$ into (3.41) yielded a similar profile to the zoomed in Figure 3.24 but created
Figure 3.24: The frequency dependence of Figure 3.23 demonstrating that the impedance is only dependent on frequency for very small values of $y/\lambda$.

A surface impedance that performed well over broad bandwidth. Using the substitution of $k'$ into (3.40) gives

$$Z(y)\bigg|_{k=k'} = \frac{j}{\omega \epsilon} \sqrt{C \omega \epsilon \alpha - \alpha^2 - \frac{1}{\lambda^2}}.$$  (3.42)

After substituting $\alpha = 0.475k'$ and $C = 1.22(k')^2$ into (3.42) we get the desired impedance profile as shown in Figure 3.25.

This surface impedance is practically realizable through different depths of corrugations, but only by a physically discretized version of Figure 3.25. To discretize the impedance, the period and wall thickness parameters of the corrugations must be known. For this, we will consider the small ground plane shown in Figure 3.9 where $\lambda_0$ corresponds to 18 GHz and therefore the
period is selected to be \( p = 0.156\lambda_0 = 2.1 \text{ mm} \) and the wall thickness is \( w = 0.03\lambda_0 = 0.5 \text{ mm} \) (both dimensions were found in [26]). Similar to the quarter wavelength corrugations shown in Figure 3.9, there is enough for room for 10 corrugations in between both waveguides. Using these parameters, the impedance in Figure 3.25 is discretized as shown in Figure 3.26. Because these corrugations are not infinite in a plane, the impedance equation for corrugations (3.10), may not be as accurate as needed for computing the depths of each of the corrugations. Instead, the input impedance of a shorted parallel plate TEM waveguide of finite width is used:

\[
Z_{in} = jZ_{0}^{PPWG} \tan(k_0d)
\]

(3.43)

where \( Z_{0}^{PPWG} = \frac{\eta_0 g}{W} \) is the TEM characteristic impedance of a parallel plate waveguide (PPWG) with plate separation distance \( g \) and plate width \( W \). The propagation constant of the PPWG, \( k_0 = \frac{2\pi}{\lambda_0} \), is multiplied by the length, \( d \), which in this case is the depth of corrugation to achieve the necessary input impedance. \( W \) for this case is simply the width of ground plane which is about 0.8\( \lambda \) at 18 GHz. Note the similarity of (3.43) to (3.1) which implies there will still be a frequency-periodic dependence of each corrugation individually; however, they will each present a different impedance (at every frequency) to the TM surface mode which can potentially force the surface
mode to radiate off of the surface at overall frequencies. Solving (3.43) for the depth yields:

\[ d = \frac{1}{k_0} \arctan \left( \frac{Z_{in}}{jZ_{0}PPWG} \right), \]

(3.44)

Plotted in Figure 3.26 is the uniformly discretized impedance from Figure 3.25 and its associated corrugation depth computed using (3.44).

![Impedance profile and corrugation depth](image)

**Figure 3.26:** The discretized impedance profile (left) and the associated corrugation depths (right) to create the tapered depth corrugations.

The depths shown in Figure 3.26 are utilized to build the corrugations on the same small ground plane model (instead of traditional quarter wavelength corrugations) as shown in Figure 3.27. These non-traditional corrugations are referred to from here on as the Tapered Depth Corrugations (TDC). The field analysis from the previous section is performed for the TDC and is shown in Figure 3.28 which compares the traditional quarter wavelength corrugations to these TDC as well at the metal only ground plane condition.

The TDC from Figure 3.28 show that the electric field is reflecting and radiating from the TDC for all four frequency increments of \( f_0 \). The reflection is evident by the strong disturbance in the contour lines such that we no longer observe smooth circular contours in the cut plane like
Figure 3.27: The small ground plane with the tapered depth corrugations computed from (3.42) and (3.44).

Figure 3.28: Cut planes of the models with the complex magnitude of the electric field plotted in dBV/m (scale on right side). The three columns represent the respective model at the top (left: Metal Only, middle: Quarter Wavelength Corrugations, and right: the Tapered Depth Corrugations), while the four rows represent frequency in integer increments of $f_0$ (18 GHz).
what is shown in the metal only condition, while the radiation is evident from the squinting of the fields close to the surface. To get a more qualitative analysis, the isolation is computed over the bandwidth and is plotted over normalized frequency in Figure 3.29 which is also compared with the quarter wavelength corrugations.

Figure 3.29: The isolation on the small ground plane plotted over normalized frequency for the tapered depth corrugations comparing it with traditional quarter wavelength corrugations.

As seen in the plot, the isolation is broadband for the tapered depth corrugations as compared to traditional quarter wavelength corrugations. In fact, the isolation improvement is consistent across at least a 4:1 bandwidth. The tapered depth corrugations are designed using a function of distance between the two antennas; whereas an antenna separation cannot be defined for traditional quarter wavelength corrugations that are designed using infinite periodicity in a plane. While the TDC are designed at a single frequency similar to quarter wavelength, it appears that they demonstrate some frequency independence and show excellent potential for use in broadband repeater systems similar to the CRS.
3.2 Isolation on a Flat Ground Plane

We have already demonstrated that corrugations in between two rectangular waveguides on a small ground plane will enhance the isolation by at least 10 dB at the frequency where the corrugations are a quarter wavelength deep. This small ground plane model is good for understanding the general impact on the coupling, but the application has a very different look than the small ground plane in Figure 3.8. A larger flat ground plane may give some more insight into how corrugations need to be designed to improve isolation on a repeater platform without requiring the computational resources of simulating the entire electrically large CRS.

3.2.1 Flat Ground Plane Numerical Studies

We consider the flat, PEC ground plane that is shown in Figure 3.30 that is 30.5 cm long by 22.9 cm wide. Two quad-ridge horn antennas that are designed to operate with consistent far-field features and impedance match over 16-45 GHz are spaced 13.7 cm apart on the ground plane to mimic a repeater platform.

For these studies, the antennas are oriented in E-plane, meaning the electric field vector in one antenna is parallel with the electric field vector of the other antenna (worst case scenario from an isolation standpoint). In the dual-polarized CRS applications, the antennas are typically oriented in the diagonal plane because the antennas are dual-polarized, which is investigated later in this chapter. Another model was constructed, but corrugations were added in between the horn antennas as shown in Figure 3.31. There are 43 corrugations that are 20.3 cm wide and extend a length of 11.3 cm between the antennas. The corrugation dimensions are the same as Figure 3.4 with a gap size of \( g = 0.126\lambda_0 = 2.1 \text{ mm} \), a wall thickness of \( w = 0.03\lambda_0 = 0.5 \text{ mm} \), and a corrugation depth of \( d = 0.25\lambda_0 = 4.1 \text{ mm} \) where \( \lambda_0 \) is the free space wavelength of the design frequency which is 18 GHz in this case.

The tapered depth corrugations were also implemented into this design as shown in Figure 3.32; however, because of the nature of the exponential increasing of (3.40), if the distance between
the two antennas becomes large then the impedance requires quarter wavelength corrugations for 
$y$ greater than approximately $2\lambda$. This is also evident in Figure 3.26 where the depth of the 
corrugations are asymptotically approaching $0.25\lambda$. The tapered depth corrugations are followed 
by a single quarter wavelength corrugation (a “Termination Corrugation” so to speak) and repeated 
periodically to fill the space between the antennas. The tapered depth corrugations are also mirrored 
about the center of the ground plane as seen from the side view in Figure 3.32 in order to maintain 
symmetry.

All three models were simulated and the isolation is computed over 15-45 GHz as shown 
in Figure 3.33. For the quarter wavelength corrugations, there is clear improvement in isolation 
starting from around 16 GHz, and a very strong improvement ($\approx 27$ dB) at the design frequency. 
However, as the frequency increases, the improvement becomes smaller similar to what was shown 
for the small ground plane in the previous section, until around 37 GHz where the isolation becomes 
worse than the ground plane alone. This is again because the corrugations are presenting an
Figure 3.31: The flat ground plane with quarter-wavelength (at 18 GHz) corrugations in between the antennas in order to improve the isolation.

Figure 3.32: The flat ground plane with the tapered depth corrugations (designed at 18 GHz) in between the antennas to improve the isolation over a wide bandwidth.

insufficient surface impedance (crossing the Smith chart from the capacitive half-plane) which is
now enhancing the propagation of the TM surface mode. The tapered depth corrugations do not have as strong of an improvement at 18 GHz, but do in fact consistently improve the isolation over the entire bandwidth as shown in Figure 3.29.

**Corrugation Density Between the Antennas**

Looking into the geometry of Figure 3.31 or Figure 3.32, the structural integrity of the ground plane may come into question as the thickness of the CRS is quarter-inch aluminum (6.35 mm) whereas the maximum depth of the corrugations are 4.1 mm. A decoupling structure such as these corrugations would leave a relatively small amount of aluminum (2.25 mm) to apply structural support for the CRS. To investigate this potential issue, the corrugations are divided into 11-corrugation sections and three additional cases (three for quarter wavelength and three for tapered depth) are considered as demonstrated by the models in Figure 3.34. The isolation for these four models plus the ground plane for both quarter wavelength and tapered depth corrugations are shown in Figure 3.35.

The results in Figure 3.35 show that as more corrugations are added, the isolation increases...
Figure 3.34: The flat ground plane with quarter-wavelength (at 18 GHz) corrugations in different increments of 11-corrugation sections. Single (top left) shows 11 corrugations next to one antenna only. Double (top right) shows 11 corrugations by both antennas (22 total corrugations). Triple (bottom left) shows three sections of corrugations, two of the sections are placed near the antennas while the third section is centered in the ground plane (33 total corrugations). Full shows the same model seen in Figure 3.31 which has 43 corrugations in between both antennas. The tapered depth corrugations have identical models, but are not shown in this thesis.

Figure 3.35: Isolation on the flat ground plane comparing the different corrugation models shown in Figure 3.34. Quarter wavelength corrugations are shown on the left and the tapered depth corrugations are shown on the right.
as well. From the standpoint of balancing both structural integrity and isolation enhancement, the “Double” case seems to make the most sense as it can still achieve a significant amount of improvement at the lower frequency without removing a significant amount of metal from the platform. Only a couple dB is gained from “Triple” to “Full” which can lead to the conclusion that once the majority of the surface has been modified, adding more will have smaller impact on the isolation.

**Concentric Corrugations**

Another consideration for repeater platforms is that it’s possible to have multiple antenna systems share a similar ground plane, and if these other systems operate in overlapping frequency ranges, the repeater’s transmitter could be detrimental to these other systems. In this case, the corrugations would need to enclose the antenna(s).

![Figure 3.36: Concentric quarter wavelength corrugations on the flat ground plane for single antenna (left) and both antennas (right).](image)

To implement this type of requirement, the corrugations can be made concentric as shown in Figure 3.36. Again the concentric corrugations have only 11 corrugations and the spacing between the antenna edge and the first corrugation is 9.82 mm. The isolation is plotted in Figure 3.37 for the two concentric cases while compared with the “Single” and “Double” cases from above for quarter wavelength corrugations. The same is done for the tapered depth corrugations in Figure 3.38. This demonstrates that the concentric corrugations have nearly the same impact as the straight corrugations and could therefore be used to minimize interference to other antenna systems on the same ground plane. There is some extra rippling in the concentric tapered depth
Figure 3.37: Isolation for the concentric quarter wavelength corrugations on the flat ground plane for single antenna (left) and both antennas (right). The plots are compared with the “Single” and “Double” cases from Figure 3.34, respectively.

Figure 3.38: Isolation for the concentric tapered depth corrugations on the flat ground plane for single antenna (left) and both antennas (right). The plots are compared with the “Single” and “Double” cases from Figure 3.34, respectively.

corrugations; however, it does not hurt the isolation as compared to the ground plane only case and still demonstrates broadband behavior as compared to the quarter wavelength corrugations.

These numerical studies have given insight into how corrugations are expected to work for isolation enhancement on a flat ground plane repeater platform. To validate these results, a flat ground plane is fabricated and measured which is covered in the next two sections.
3.2.2 Flat Ground Plane Fabrication and Assembly

The results presented in the previous sections demonstrated the theory, design, and isolation expected from placing corrugation in between two antennas on the same ground plane.

To validate these principles, a flat ground plane was fabricated and is shown in Figure 3.42 and the fabricated dimensions in Figure 3.39. The ground plane was fabricated in such a way

Figure 3.39: Dimensions of the 3D model of the fabricated flat ground plane showing the holes for the antennas, decoupling structures (i.e. corrugations), and flush mounted screws. All dimensions shown are in units of centimeters

Figure 3.40: Quad ridge horn antennas designed in [103] as well the custom 3D printed holder to mount the antennas to the fabricated flat ground plane
to accommodate different decoupling structures which could be brought from the bottom side and attached using flush mountable screws from the top side as demonstrated in Figure 3.42. The horn antennas are connected in a similar fashion using a custom 3D printed holder (shown in Figure 3.40). Copper tape was then used to cover the seams as well as to cover up the corrugations to create a metal only condition as seen in Figure 3.43b. The ground plane and the corrugations (Figure 3.41) were fabricated using CNC machining with a gap of 2.1 mm with a wall thickness of 0.5 mm and a depth of 4.1 mm ($0.25\lambda_0$ at 18 GHz). The fully assembled fabricated ground plane with corrugations is shown in Figure 3.43a.

Figure 3.41: (a) Fabricated quarter wavelength (at 18 GHz) corrugations. (b) Fabricated tapered depth corrugations based on Tereshin’s Surface Impedance Equation (3.29)
Figure 3.42: Assembly procedure for the fabricated flat ground plane designed to accommodate different decoupling structures.
Figure 3.43: (a) Fully assembled fabricated ground plane with corrugations. Copper tape is used to seal the small gaps between the antenna and corrugations and the ground plane. (b) Fully assembled fabricated ground plane with copper tape placed over the corrugations to create a “Metal Only” condition.

3.2.3 Flat Ground Plane Measurements

Due to the high gain of the antennas at boresight, measurements indoors will become distorted due to reflections from the ceiling. Even inside an anechoic chamber the small reflections may still distort the isolation. There are two ways to overcome this: use time domain analysis to perform time gating on the measured signal to determine the true coupling (see Appendix A) or to take measurements outdoors where the antennas can radiate into free space. In this section, the measurements were performed outdoors with the measurement setup shown in Figure 3.44. The precision network analyzer (PNA) is set to measure 801 frequency points from 15-40 GHz. Due to
the hardware limitations of the PNA, 40 GHz is the highest frequency measurable. The Intermediate Frequency Bandwidth (IFBW) is set to 20 Hz to ensure the noise floor (see Appendix B) is low and the output power is set to 0 dBm.

Figure 3.44: Measurement setup for the fabricated flat ground plane. To minimize the impact of scatterers (such as the ceiling), the measurements were taken outdoors using a cart. On the top of the cart is the device under test, while beneath the cart is the network analyzer as shown.

With the measurement setup and the fabrication design, measurements are taken for both the tapered depth corrugations and traditional quarter wavelength corrugations for both double and single cases similar to Figure 3.34. As shown in Figure 3.45, the measured isolation demonstrates the bandwidth limitation of the quarter wavelength corrugations, while also validates the broadband behavior of the tapered depth corrugations. There is also a significant improvement in performance in going from single to double corrugations for both quarter wavelength and tapered depth corrugations, which also validates the previous numerical results.
Figure 3.45: Isolation for the fabricated flat ground plane comparing quarter wavelength corrugations, tapered depth corrugations, and metal only cases for single (left) and double (right) sets of corrugations. The picture above each plot demonstrates the implementation of the single and double cases.

3.3 Isolation on a Cylinder Ground Plane

In the previous sections, two approaches to corrugations were explained in detail both theoretically and through numerical studies and further validated through fabrication and measurements on a flat ground plane. Quarter wavelength corrugations are a well known technique that works very well over a limited bandwidth. From Tereshin’s surface impedance equation (3.29), the tapered depth corrugations were investigated and a practical broadband decoupling structure was developed and validated on the flat ground plane. This section looks to implement the corrugations onto the CRS to satisfy the application requirements for isolation and demonstrate feasibility of such corrugations on a cylindrical ground plane.
3.3.1 Numerical Studies

To begin these numerical studies, a practical model of the cylinder needs is found because simulation of the full CRS as a 15.2 cm diameter, 30.5 cm long cylinder with two orthomode transducer’s (OMT’s) exciting both antennas is a computationally extensive problem. To start, the OMT’s are not used in these simulations to simplify the model and instead the quad ridge horns are excited with a double ridge waveguide and short dual-to-quad ridge transition. It is safe to assume that shrinking the length of the cylinder should result in a significant change the magnitude of the coupling. The separation between the antennas (\( R \)) will decrease and therefore the isolation will decrease, but through approximation using Friis transmission formula (1.1) we can expect the isolation to decrease roughly as \( \frac{1}{R^2} \). Therefore, the model used is only 18 cm long instead of 30.5 cm, and the antenna separation is 11.9 cm instead of 20.3 cm. It can be then expected that the actual isolation will be scaled by \( \left( \frac{20.3 \text{ cm}}{11.9 \text{ cm}} \right)^2 \) or in other words, the actual isolation of the CRS should be about 4.6 dB higher than the simulation results presented here. Modifying the length may also have additional impact on the trend of the curve as standing waves on the surface of the cylinder will be dependent on the length; however, these impacts are considered secondary to the overall magnitude of the isolation. The 15.2 cm diameter is maintained to ensure the antenna aperture is approximately the same as on the CRS. For these initial experiments, the antennas are oriented in the E-Plane with respect to each other.

Three models are first considered and are shown in Figure 3.46 including the finite element boundary integral (FEBI) box that encloses each of the models. To simplify the volume of the mesh, the inside surface of the cylinders are also set to an FEBI boundary condition. All three models were simulated and the isolation is presented in Figure 3.47, which shows the difference in isolation between the different models is acceptable for these studies. The computation time and RAM required per frequency point for the three cylinder ground plane models is presented in Table 3.1 demonstrating the efficiency of using a smaller model to approximately compute the same magnitude of isolation.
Figure 3.46: The three models considered for efficient simulation of the CRS. The bottom shows a full 15.2 cm diameter, 30.5 cm long cylinder, the top left shows the same cylinder but only keeping the top half, and the top right shows a 45° cut of the same cylinder.

Figure 3.47: The isolation for the three models in Figure 3.46.
Table 3.1: Computation time and RAM required per frequency point for the three cylinder ground plane models.

<table>
<thead>
<tr>
<th>Model</th>
<th>Solve Time [minutes]</th>
<th>RAM [GB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Full Cylinder</td>
<td>200</td>
<td>105</td>
</tr>
<tr>
<td>Half Cylinder</td>
<td>39</td>
<td>34.2</td>
</tr>
<tr>
<td>45° Cylinder</td>
<td>11</td>
<td>7.45</td>
</tr>
</tbody>
</table>

E-plane Orientation

Both traditional quarter wavelength corrugations and the tapered depth corrugations are implemented onto the 45° cylinder ground plane as shown in Figure 3.49 and the isolation is reported in Figure 3.48. The results are as expected and agree very well with the flat ground plane, showing the quarter wavelength corrugations have a limited 2:1 bandwidth and degrade the isolation performance at around double the design frequency whereas the tapered depth corrugations demonstrate consistent isolation enhancement across a broad bandwidth.

Figure 3.48: The isolation of the traditional quarter wavelength corrugations and the tapered depth corrugations shown in Figure 3.49 also compared with the 45° metal cylinder only.
Figure 3.49: (a) Quarter wavelength (at 18 GHz) corrugations and (b) tapered depth corrugations implemented onto the 45° cylinder ground plane.

**D-plane Orientation**

The good agreement between flat ground plane and the cylinder ground demonstrates that for antennas oriented in E-plane, the tapered depth corrugations will provide consistent isolation enhancement for over a 2.5:1 bandwidth. However, the application of the CRS requires the antennas be dual polarized and in a 45° orientation or diagonal plane (D-plane). This can be easily adjusted in the model of the 45° cylinder and is done so. By orienting the antennas in D-plane, there are now two possible paths of coupling due to the two orthogonal polarizations of the antennas demonstrated by Figure 3.50. There could be coupling from polarization (Pol-) 1 to Pol-1 or from Pol-1 to Pol-2, and therefore the two coupling paths should be distinguished. The isolation is plotted in Figure 3.51 for the cylinder only case to show how the isolation may be different between
these two polarizations and additionally, the E-plane isolation from the cylinder only case is also shown for comparison.

Figure 3.50: A top down view of the 45° to visualize the different antenna polarizations in the D-plane.

Figure 3.51: The isolation for the D-plane cylinder only model showing how the isolation may be different between these two polarizations. Additionally, the E-plane isolation of the cylinder only from Figure 3.48 is also shown for comparison.

For the TM surface mode propagating along the y-axis between the two antennas, both Pol’s will have common field components, therefore, it is expected that the Pol-1 to Pol-1 isolation should
be nearly identical to the Pol-2 to Pol-2 isolation. If the model was perfectly symmetric including the numerical mesh, it is expected that these curves would be line up on top of each other. This also concludes that only half the numerical calculations are required since both levels of isolation are identical. Also note that the isolation has improved compared to E-plane which is also expected due to the concentration of the fields at the aperture is no longer in the direct path between the antennas.

After orienting the antennas in the D-plane, the isolation is computed for the quarter wavelength and tapered depth corrugations and compared with the cylinder only case shown in Figure 3.52. There was a sharp ripple seen in the isolation for the tapered depth corrugations as compared to the E-plane orientation case at lower frequency (specifically around 21 GHz). This behavior was also found in several other models (both flat and cylindrical ground planes) that were investigated and the effect was studied by utilizing the surface currents at these frequencies.

Figure 3.52: The isolation of the traditional quarter wavelength corrugations and the tapered depth corrugations compared with the cylinder only case when the antennas are oriented in the D-plane. The sharp ripple in the tapered depth corrugations is circled in the image.

**Improving Low Frequency Performance**

At low frequency the length from the first corrugation to the last corrugation is not considered electrically large (at 18 GHz, this length is $1.56\lambda$) which allows for standing waves along the length
of the corrugations, for instances when the total magnetic field is not parallel with the corrugations. This creates a hard surface effect for this wave along the length of the corrugations which results in some leakage of the D-plane wave across to the other antenna.

Figure 3.53: Slots cut into the corrugations to create a “domino” look.

In order to improve this sharp ripple at lower frequency to maintain consistent isolation enhancement, the hard surface effect needed to be prevented, which was done by segregating or cutting up the corrugations as shown in Figure 3.53. The slots are spaced about $0.8\lambda_0$ at 18 GHz which is roughly the same length as when the tapered depth corrugations were designed using the parallel plate waveguide input impedance equation (3.43). These slots prevent the wave propagating along the corrugations from seeing a zero longitudinal impedance which prevents a type of hard surface condition, but they do not degrade the impact of the wave propagating along the path between the two antennas as shown in Figure 3.54. The results show the ripple at low frequency
has been suppressed and the broadband performance of the tapered depth corrugations are still present. The quarter wavelength corrugations were impacted only slightly, but the bandwidth limitation is not impacted at all.

![Figure 3.54: Isolation after adding the slots to create the “domino” look with the corrugations to help performance at lower frequency.](image)

### 3.3.2 Cylinder Ground Plane Fabrication and Assembly

The CRS with the tapered depth profile was then fabricated using 3D printing and copper plating techniques. An aluminum tube with the same dimensions as the CRS was cut into two sections that would hold both antennas, while the tapered depth corrugations were printed using a stereolithography (SLA) 3D printing technique in four parts that would connect together to form a cylinder as shown in Figure 3.55a. The parts were then copper plated (Figure 3.55b) in order to make the corrugations metallic. About 35.56 µm (1.4 mil) of copper was added which should be more than adequate in the 15-45 GHz range as the skin depth of copper at 10 GHz is about 1 µm (0.04 mil). The plated parts are then integrated with the aluminum cylinder parts as shown in the assembly in Figure 3.56. Copper tape was used to seal the aluminum cylinder parts to the 3D printed parts as well as to seal the inside of the cylinder to prevent leakage in between any of the parts from interfering with the isolation measurements.
Figure 3.55: 3D printed tapered depth corrugations implemented onto a cylinder in four parts shown (a) before copper plating and (b) after copper plating.

Figure 3.56: Assembly of the tapered depth corrugations implemented on the CRS.
3.3.3 Cylinder Ground Plane Measurements

The same measurement setup presented in Figure 3.44 (however it was on a different day) was used for measuring the CRS with the tapered depth corrugations: measured outside, 15-40 GHz measurement range, 20 Hz IFBW, 801 points, and 0 dBm power. To create the “cylinder only” condition (an ideally solid metal cylinder with no corrugations), copper tape was used to cover, or short out, the tapered depth corrugations (Figure 3.57) similar to what was presented for the flat ground plane. This technique was validated by comparing the original measured CRS isolation (see Figure 1.4) to the cylinder only measurement performed here.

Figure 3.57: Copper tape applied over the tapered depth corrugations to short the corrugations in order to create a cylinder only condition.

![Copper tape applied over the corrugations](image)

Figure 3.58: Measured isolation of the CRS with and without the tapered depth corrugations. Copper tape was applied over the corrugations to create the cylinder only condition.

![Measured isolation graph](image)
The measured isolation with and without the tapered depth corrugations is shown in Figure 3.58. As expected from the numerical results, the tapered depth corrugations provide broadband isolation enhancement. Traditional quarter wavelength corrugations were not fabricated, but a simulation model was generated using the shorter cylinder model from Figure 3.46 and is shown along with a model for the tapered depth corrugations in Figure 3.59.

The simulated isolation is compared in Figure 3.60 which shows the same trend seen throughout this chapter. The general trend between the measurements and simulations have good agreement; however, it would not be a fair comparison to plot them against each other for a few reasons. First, the measurements are with an OMT which was not modeled due to computational resources which small reflections (due to mismatch losses) and physical imbalances may have a very strong impact on the high isolation levels. Second, the cylinders are not the same length, which as stated earlier, will account for about 5 dB difference in overall magnitude. Third, the antennas are more inset into the cylinder than the simplified model. Finally, the copper tape and fabrication tolerances make the surface of the cylinder wobbly which would become extremely difficult to model accurately especially when considering isolation levels of 70 dB. However, a fair figure of merit to compare the simulations and the measurements is the isolation improvement shown in Figure 3.61 which is the isolation of the corrugations normalized to the isolation of the cylinder only. In general, the trend agrees very well between the simulated and measured isolation improvement. The discrepancy is again attributed to differences between the physical and numerical models, but the simulation gives good insight into how the tapered depth corrugations are performing over traditional quarter wavelength corrugations. Moreover, the simulation results of the quarter wavelength corrugations are validated by the measurements performed on the flat ground plane.
Figure 3.59: Full short cylinder (18 cm long) models of the CRS (left), quarter wavelength corrugations (middle), and tapered depth corrugations (right).

Figure 3.60: Simulated isolation of the models shown in Figure 3.59.

Figure 3.61: Comparing the simulated and measured isolation improvement for the CRS with corrugations.
As eluded to earlier, the corrugations present a potential issue to the structural integrity of the CRS. Even the tapered depth corrugations would require a very thin amount of metal remaining at certain points along the cylinder. To try to mitigate this issue, a coverage angle ($\phi$) is defined as the angle over the cylinder that the corrugations cover. This coverage angle was investigated both numerically and experimentally. This angle describes how much of the cylinder needs to be converted into corrugations to achieve the adequate amount of isolation improvement. The coverage angle was physically implemented by applying copper tape in increments along the cylinder as shown in Figure 3.62.

A larger coverage angle corresponds to requiring more of the cylinder to be machined off, therefore, these measurements sought to minimize $\phi$ yet maintain the full isolation improvement possible from the tapered depth corrugations. The measurement results are plotted in Figure 3.63. As seen, there were 12 $\phi$ measurements taken in total not including the “Cylinder Only” which
was used as a baseline. For easy viewing of the multiple curves shown on each plot, a smoothing function [104], is used to help visualize all the measured values of $\phi$. For even easier visualization, the isolation improvement is averaged over all measured frequency points from 18-40 GHz and plotted versus $\phi$ as well. These results demonstrate that for $\phi > 60^\circ$, the average isolation improvement saturates and therefore it can be concluded that the entire cylinder would not need the corrugations all the way around, but only directly between the two antennas within this $60^\circ$ angle. This also qualitatively validates the numerical studies in Figure 3.47 as the majority of isolation comes from the direct path between the two antennas and an insignificant portion of the isolation is attributed to surface currents wrapping around the cylinder. Additionally, there was an outlier around $\phi = 35.2^\circ$ which had significant improvement between 20-25 GHz which made the average jump significantly higher than the other measurements. These results were verified numerically as shown in Figure 3.64. Note these observations may not necessarily be general for any cylinder, but similar conclusions could be easily applied to cylinders with diameters much larger than a wavelength (see Chapter 2).

Figure 3.63: The measured isolation (after application of smoothing function) for all measured values of $\phi$ plotted over frequency (top). The isolation improvement averaged over all measured frequency points from 18-40 GHz plotted over $\phi$ (bottom). The outlier at $\phi = 35.2^\circ$ with higher isolation improvement is also shown.
Figure 3.64: Some of the simulated models to mimic the coverage angle measurements. A 120° cylinder matching the simulated models from above was used (left). In total, simulation 9 points were taken in order to validate the $\phi > 60^\circ$ saturation point from the measurements. The isolation improvement averaged over all simulated frequency points from 18-45 GHz plotted over $\phi$ (right).

3.4 Summary

This chapter presented corrugations as an impedance surface for isolation improvement in wide extents. Theoretical concepts from Chapter 2 were extended to apply to the specific case of corrugations and a close form expression describing the surface reactances was derived based on the Fresnel reflection equations [6,78], and comparison with full-wave simulators demonstrated the accuracy of the tool used throughout these studies. Uniform depth corrugations have a bandwidth limitation that is fundamental to its geometry, which was overcome by investigating techniques such as Tereshin’s Surface Impedance Equation (3.29) [25, 26]. Utilizing this expression, tapering the depth of corrugations could allow for the surface impedance to perform more broadband in terms of isolation improvement. This was tested numerically and experimentally and wide bandwidth performance was validated. Practical implementation onto flat and cylinder ground planes for millimeter wave repeater applications was demonstrated and a path has been paved for future theoretical development on the tapered depth corrugations.
4.1 Introduction to Reactive Impedance Surfaces

The reactive impedance surfaces (RIS's) discussed in this chapter demonstrate similar behavior to corrugations, but are implemented differently. Because the geometry of such impedance surfaces are significantly more complex, there is no closed form derivation for the surface impedance like the tangent function for the corrugations. However, typically these surfaces exhibit a surface impedance that can be approximated as:

\[ Z_s \approx j [\tan(k\delta) + X_{L,C}(\omega)] \]  \hspace{1cm} (4.1)

where \( k \) is the material wavenumber, \( \delta \) is related to some dimension of the RIS (usually the material thickness and is frequency thickness), and \( X_{L,C}(\omega) \) is extra added frequency dependent reactance (usually implemented in inductive (\( L \)) or capacitive (\( C \)) loading). These parameters are not typically calculated or used in the design process, but (4.1) can give insight into how the surface impedance acts. RIS's performs similar to corrugations through the periodic behavior of the tangent function; however, the additional loading will change the bandwidth performance of the surface. When the tangent function goes to zero, if \( X_{L,C}(\omega) < 0 \), the surface will still be capacitive because \( Z_s < 0 \). From the standpoint of preventing the TM mode from propagating (which would require a capacitive surface impedance), this loading will extend the capacitive bandwidth of the surface and therefore create a more broadband impedance surface.
4.1.1 Loaded Corrugations Example

To visualize this bandwidth improvement, we consider the modified corrugations in Figure 4.1 which will be referred to as “loaded corrugations” due to the top loading. Also shown is the unit cell with its parameters that can be defined for this geometry: $d$ is the depth of the corrugation, $p$ is the period, $w$ is the wall thickness, $h$ is the width of the PEC “hat” that has been added, and $t$ is the thickness of the hat. The addition of a hat is a well known capacitive loading technique for antenna miniaturization due to the hat creating a type of parallel plate capacitor with the ground [105–108].

![Figure 4.1: Illustration of the loaded corrugations including the parameters of the unit cell. Loaded corrugations are essentially regular corrugations with the addition of a “hat” which capacitively loads the corrugations.](image)

Using the unit cell in Figure 4.1, the Floquet analysis is performed on these loaded corrugations to find the surface impedance. Similar to the corrugations studied in the previous chapter: $d = 0.25\lambda_0$, $w = 0.03\lambda_0$, and $p = 0.156\lambda_0$; the hat loading had an arbitrarily selected thickness of $t = 0.01\lambda_0$ and the hat width was varied to change the $X_{L,C}$ of this impedance surface. The computed TM surface reactance is shown for an incidence angle of 80° in Figure 4.2. Case $h = 0$ (no loading is added) is the equivalent of standard quarter wavelength corrugations. When this loading is added, the parallel resonant frequency shifts below $f_0$ as $h$ increases in size; however, the series
Figure 4.2: The TM surface reactance of the loaded corrugations plotted over normalized frequency where \( f_0 \) is the design frequency. The different curves represent the parametric analysis of the hat width, \( h \) for the rest of the parameters from Figure 4.1 held constant.

Table 4.1: Relative bandwidths for the loaded corrugations. The quantity, \( f_{\text{min}} \), refers to the first parallel resonance of the surface reactance while the bandwidth is the ratio of \( f_{\text{max}} \) (the maximum frequency where \( X_{\text{TM}} < 0 \)) to \( f_{\text{min}} \).

<table>
<thead>
<tr>
<th>( h )</th>
<th>( f_{\text{min}} / f_0 )</th>
<th>Bandwidth ( [f_{\text{max}} : f_{\text{min}}] )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>1</td>
<td>2.1:1</td>
</tr>
<tr>
<td>( 0.05\lambda_0 )</td>
<td>0.87</td>
<td>2.3:1</td>
</tr>
<tr>
<td>( 0.1\lambda_0 )</td>
<td>0.75</td>
<td>2.7:1</td>
</tr>
<tr>
<td>( 0.14\lambda_0 )</td>
<td>0.5</td>
<td>4:1</td>
</tr>
</tbody>
</table>

resonant frequency (the second zero-crossing point) does not shift with the only exception of \( h = 0 \), which is a result of increasing the depth of the other loaded corrugation by thickness \( t \). However, this shift is insignificant relative to the shifts occurring below the parallel resonance which demonstrates larger capacitive bandwidth of the corrugations. The specific relative capacitive bandwidths for this study are shown in Table 4.1.

**Scaled Loaded Corrugations**

As seen above, loaded corrugations have wider capacitive bandwidth than traditional corrug-
gations and therefore prevent the TM mode from propagating over a wider range of frequency. In order to achieve this behavior over the desired bandwidth, the corrugation depth would need to be scaled to shift the parallel resonance back to $f_0$. For example, consider the case where $h = 0.1\lambda_0$ where the parallel resonance occurs roughly around $0.75f_0$, indicative that the corrugations are deeper by a factor of $\frac{4}{3}$. To fairly compare the bandwidht of this case with traditional quarter wavelength corrugations ($d = 0.25\lambda_0$), the depth of the loaded corrugation is shrunk to $d = 0.19\lambda_0$ and the results are given in Figure 4.3. This scaled loaded corrugation now has the same parallel resonant frequency as the traditional quarter wavelength corrugation, but the capacitive loading improves the capacitive bandwidth of the loaded corrugations to be at least 2.5:1. Therefore, this loading will allow for the suppression of the TM surface mode over a wider bandwidth than the traditional quarter wavelength corrugations. In relation to (4.1), these loaded corrugations have a tangent function behavior from which we can say that $\delta$ and $X_{L,C}$ are functions of $d$, $h$, and $t$. However, finding a closed form expressions for these functions was not pursued as part of this thesis. Typically, these functions are solved for using equivalent circuit models of the surface [18, 40, 109].

Figure 4.3: The TM surface reactance of the scaled loaded corrugations plotted with the TM surface reactance of a traditional quarter wavelength corrugations in order to compare the bandwidth improvement with loaded corrugations.
demonstrated in earlier work [91]; however, it is used here to introduce the concept of a broadband RIS and how it can achieve wider than the 2:1 bandwidth associated with traditional corrugations.

### 4.1.2 Evolution from Loaded Corrugations to Other RIS’s

Consider the geometry of the loaded corrugations in Figure 4.1 for a wave that is incident parallel to the corrugations, the surface behaves as a PEC sheet. This behavior is undesirable. For an unknown incidence angle, the geometry of the surface should be designed to be omnidirectional. Figure 4.4 presents a simplified evolution to make the surface omnidirectional where the hat and walls are broken up along the corrugations to have the same period in both directions. The broken hats and walls are converted to cylinders which creates a nearly omnidirectional impedance surface. Square hats present a surface impedance that is close enough to omnidirectional that they are widely implemented throughout literature. Finally, if the cylinder hats are removed (or if the same evolution is applied to corrugations without loading) the resulting structures will be an array of cylinders. This exactly resembles that of a bed of nails or bed of pins (see Figure 2.5), which has an identical TM surface impedance to corrugations (3.1), but independent of azimuth angle.

![Figure 4.4: The illustration of the evolution steps from the loaded corrugations to a double periodic reactive impedance surface (RIS).](image)

Alternatively, each corrugation/pin row could be machined separately and then arranged one after the other to create the surface. The corrugations/pins could also be machined and the hats fabricated separately and electrically attached to the corrugations/pins. Another fabrication
technique would be through 3D printing similar to the process outlined in the previous chapter for
the tapered depth corrugations on the cylinder. A fabrication technique which is typically used for
these types of surfaces involves etching and plating printed circuit boards (PCB’s). This technique
is widely used, low cost, and well developed. In a PCB, the unit cell of these surfaces tends to
mimic two copper layers with a via connecting them. A substrate which fills the gap between the
bottom and top layers of the surface adds capacitance because the dielectric constant inside the
surface has changed. As a final thought, Figure 4.4 has only shown flat surfaces. Implementation
on a cylinder should be considered as the surface geometry could limit how the surface is fabricated.
PCB’s are used in these studies because practical, soft substrates exist that can be bent around
the considered Cylinder Repeater System (CRS) with radius of curvature complying with $kR \gg 1$.

4.2 Connected Patches RIS

The concept of using a RIS to improve the isolation is not new, so the initial investigations
began with a literature search that was presented in Chapter 1. In [73], Wu et al. designed a RIS to
achieve a capacitive TM surface impedance and nearly zero TE surface impedance for greater than
a 2.5:1 bandwidth. Their work was used as a model herein was scaled to the frequency range of the
CRS. The geometry of the RIS adopted in these studies is shown in Figure 4.5 which is essentially
a double periodic (with period $p$) array of rectangular patches ($l \times w$) that are connected together
with small metallic strips (width $s$) and shorted to a ground plane with a via (on a substrate of
thickness $t$).

These studies begin with investigating the rectangular patches by themselves (i.e. let $s = 0$)
on an air dielectric ($\varepsilon_R = 1$). To synthesize the RIS, the design frequency, $f_0$, is set as the lowest
frequency in the bandwidth of interest. Therefore, for the surface to have capacitive behavior
across the bandwidth, the first parallel resonance should be located at $f_0$. Based on the design
in [73], the period of the unit cell is chosen to be $p = 0.1\lambda_0$, the thickness of the air substrate is
t = 0.174$\lambda_0$, the diameter of the via is 0.01$\lambda_0$, and the width of the patch is $w = 0.067\lambda_0$ where
$\lambda_0$ is the freespace wavelength at $f_0$. The length of the patch ($l$) is studied parametrically and
Figure 4.5: The geometry of the connected patches RIS including the parameters of the unit cell.

presented here; however, in practice all of the parameters need to be varied to achieve the optimal dimensions of the unit cell [73]. Due to the unit cell only being partially filled with metal (unlike corrugations), the TE surface mode may become an isolation issue if excited. In other words, if the surface presents a capacitive surface reactance to the TE surface mode, the RIS can potentially be detrimental to the isolation even if the TM surface mode is suppressed. Therefore, both the TM and TE surface reactances are plotted in Figure 4.6 for the different selected values of \( l \) computed using the Floquet analysis for an 80° incident angle.

Figure 4.6: TM and TE surface reactances of the rectangular patch RIS for different values of patch length, \( l \). The geometry is shown on the left.
In order to have a capacitive TM surface reactance over at least a 2.5:1 bandwidth starting from $f_0$, then the length of the patches should be $l \geq 0.08\lambda_0$. Additionally, the TE surface reactance is inductive over the same 2.5:1 bandwidth which shows that the TE surface mode would not be excited for this surface independent of $l$. However, it is also possible to eliminate this issue altogether by shorting the patches together along the transverse direction as originally shown in Figure 4.5. By placing continuous metal along the transverse direction, the tangential electric field must be zero across the unit cell which will force the TE surface mode to vanish on this RIS. This effectively turns the RIS patches into a form of loaded corrugations because corrugations will not allow the TE surface mode to be present. Adding these strips can also be thought of as creating small inductors between the unit cells which forces the TE surface mode to see a low impedance inductive surface similar to the surface impedance of a good conductor. These thru strips are added with a strip width of $s = 0.015\lambda_0$ and the surface reactances are shown in Figure 4.7 for the same parametric analysis.

![Figure 4.7: TM and TE surface reactances of the connected patches RIS for different values of patch length, $l$, demonstrating the thru strips have no impact on the TM surface reactance but short out the TE surface reactance. The geometry is shown on the left.](image)

Comparing Figures 4.6 and 4.7 shows that the TM surface reactance is almost entirely independent of these shorting strips, but the TE surface reactance is greatly impacted with small magnitude thus performing as expected. Adding these strips is also desired because it produces a soft surface over a large bandwidth because the TE mode sees a very small impedance while the
TM mode is presented with a large impedance \cite{73}. Observed soft surface behavior motivated the further maturity of this design to be used on the CRS using a patch length of \( l = 0.08\lambda_0 \) as the final design parameter.

**Practical Fabrication of the Connected Patches RIS**

From a practical standpoint, the fabrication of such a surface on an air dielectric as the one shown in Figure 4.5 at frequency \( f_0 = 18 \text{ GHz} \) (\( \lambda_0 = 16.67 \text{ mm} \)) would either require extremely tight tolerances on a 3D printer or “mass production” of rows using conventional machining techniques. So, implementation onto a PCB must be considered and therefore a physical substrate should be used in the design of this RIS. To avoid adding too much capacitance to the surface, a low dielectric constant is desired, as numerical studies show that high dielectric constants can limit the bandwidth of such structures. A Rogers 5880 RT/Duroid substrate is chosen because its low loss (\( \tan \delta = 0.0009 \)) and satisfies this requirement (\( \varepsilon_R = 2.2 \)) \cite{110}. Naturally, as this substrate is added, the additional capacitance of the dielectric will impact the dimensions of the RIS in order to achieve the same surface reactance properties. However, these dimensions can be scaled by the square root of the first order approximation of the effective dielectric constant of a microstrip line \cite{23}:

\[
SF = \sqrt{\frac{2}{1 + \varepsilon_R}} \approx \frac{1}{\sqrt{\varepsilon_{\text{effective}}}}
\]  

(4.2)

where the scale factor, \( SF \), multiplies the unloaded dimension to achieve the scaled typology. While these substrates can come in any thickness, it is often more time convenient and cost effective to use standard values. Considering the case of the CRS, to minimize the profile of the RIS on the cylinder, a 1.575 mm (62 mil) substrate is chosen. After selecting a standard thickness substrate, the dimensions of the unit cell need to be reconsidered in order to match a fixed \( t \). The unit cell dimensions were studied parametrically with the real substrate to shift the parallel resonant frequency back to \( f_0 \) taking into consideration the tolerances and fabrication limitations of PCB manufacturing. Table 4.2 lists the dimensions of these three cases of the RIS: air dielectric, scaled using (4.2) on Rogers 5880, and the practical fabrication dimensions. The TM reactance of the
Table 4.2: Dimensions of the connected patches RIS for different dielectric cases. All length dimensions have been normalized to the free space wavelength $\lambda_0$.

<table>
<thead>
<tr>
<th>RIS Model</th>
<th>Periodicity $p$</th>
<th>Patch Width $w$</th>
<th>Patch Length $l$</th>
<th>Strip Width $s$</th>
<th>Substrate Thickness $t$</th>
<th>Relative Permittivity $\varepsilon_R$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Air Substrate</td>
<td>0.100</td>
<td>0.067</td>
<td>0.077</td>
<td>0.015</td>
<td>0.174</td>
<td>1</td>
</tr>
<tr>
<td>Scaled on Rogers 5880 using (4.2)</td>
<td>0.079</td>
<td>0.053</td>
<td>0.061</td>
<td>0.010</td>
<td>0.137</td>
<td>2.2</td>
</tr>
<tr>
<td>Fabricated on Rogers 5880</td>
<td>0.100</td>
<td>0.067</td>
<td>0.090</td>
<td>0.015</td>
<td>0.094</td>
<td>2.2</td>
</tr>
</tbody>
</table>

Figure 4.8: TM surface reactance of the connected patches RIS comparing the air substrate to the scaled RIS using (4.2) as well as the fabricated RIS. The dimensions of the unit cell for each case are presented in Table 4.2.

The fabricated dimensions show significant difference from simple scaling demonstrating the impact of the substrate thickness $t$. The impact of adding a higher dielectric constant is also seen as the curve is “pulled” closer to zero for the scaled and fabricated surface reactances. The higher the dielectric constant of the substrate, the tighter this curve is pulled towards zero, which is why the low dielectric constant of Rogers 5880 is desirable. The fabricated RIS shows it can achieve
a capacitive bandwidth of almost 3:1, however around $2f_0$, the surface should not be considered high impedance due to the small magnitude (significantly smaller than $\eta_0 \approx 377 \, \Omega$) of the surface reactance. The connected patches RIS, it built onto a small ground plane as shown in Figure 4.9 in an $14 \times 8$ array.

Figure 4.9: Small ground plane with $14 \times 8$ array of the connected patches RIS implemented to improve the isolation.

The isolation with the RIS is computed and presented in Figure 4.10 demonstrating significant isolation enhancement over a 2.5:1 bandwidth. If compared with the surface reactance of Figure 4.8, there is good agreement between the capacitive bandwidth of the RIS and the bandwidth of isolation enhancement, therefore validating the surface impedance modeling approach of these types of impedance surfaces. Now, more rigorous models can be investigated including the impact of integrating with the CRS.
Figure 4.10: Isolation of the connected patches RIS on the small ground plane compared with the metal only case demonstrating the bandwidth of the RIS to improve isolation.

4.3 Proof of Concept Testing with the Connected Patches RIS

4.3.1 Numerical Studies on a Small Cylinder

After demonstrating good performance in the Floquet analysis and the small ground plane, the connected patches RIS is considered for use on the CRS. The studies started with the investigation of the air dielectric typology implemented onto the cylinder in Figure 4.11 which has diameter $D_c = 38.1$ mm (a quarter of the CRS cylinder), length $L_c = 90$ mm, and antenna spacing of $S_c = 75$ mm. To simplify the required computational resources, instead of simulating the horn antennas, quarter wavelength monopoles (at each frequency point) are used.

With a small cylinder as a baseline model, the air dielectric RIS is added in different array sizes to compute how much of the RIS may be required for improving the isolation. These different RIS arrays are shown in Figure 4.12. Note that the $0 \times 0$ array is the same as the “Cylinder Only” case shown in Figure 4.11. While the RIS’s dimensions were kept fixed based on $f_0 = 18$ GHz, resonant quarter wavelength monopoles were tested at three frequencies: $f = 18, 31.5, \text{ and } 45$ GHz and the isolation is computed using Method Of Moments solver, FEKO [111]. The results for the different array sizes are compared in Figure 4.13 showing that a larger array size will result in
Figure 4.11: FEKO model of the small cylinder with parameters used for testing the isolation of the connected patches RIS with quarter wavelength monopoles the TX and RX antennas.

Figure 4.12: FEKO models of the small cylinder with different sizes of the connected patches RIS’s.

higher isolation. To help visualize this, the normalized surface current density at $f = 18$ GHz for the different RIS arrays are shown in Figure 4.14.

The surface currents show that the RIS prevents the current from propagating along the cylinder between the antennas. Additionally, a standing wave can be seen in between the RIS and transmit antenna demonstrating the current is reflecting off the surface which is the behavior
Figure 4.13: Isolation on the small cylinder computed at three different frequency points within the CRS’s bandwidth for the different RIS array sizes. The $0 \times 0$ array is the baseline (i.e. no RIS).

Figure 4.14: Normalized surface current density plotted in dBA/m for the different RIS array sizes at 18 GHz demonstrating that as the array size increases, the surface mode is more suppressed between the two antennas. The transmitting monopole is located at the top and the receiver monopole is at the bottom of each plot.
expected from a high impedance surface. The isolation plot in Figure 4.13 shows that for the $21 \times 21$ array size, 17 dB of isolation enhancement is achieved at the design frequency while at 45 GHz, 12 dB of improvement is obtained. This further validates the capacitive bandwidth computed in the Floquet analysis in Figure 4.8.

Even though these three frequency points appear to demonstrate isolation improvement over the entire bandwidth of interest, the isolation for the $21 \times 21$ array was also computed across the band in 1 GHz increments (still utilizing a quarter wavelength monopole at each frequency point). To further validate this, the same model was also computed in HFSS and the results are compared in Figure 4.15. While there are discrepancies between the two codes, in general, the same trend is seen for both methods. The disagreement could be attributed to the mismatch in the antennas themselves. In FEKO, a wire port is used to create the monopole whereas in HFSS, a lumped port is used. Slight differences in mismatch loss in the antennas between the models can influence

![Figure 4.15: Isolation comparison between FEKO and HFSS for the small cylinder with and without the $21 \times 21$ RIS array.](image)

the overall magnitude of the isolation as small reflections (on the order of 20 dB) are not small when compared with 50 dB of isolation. Additionally, the mesh plays a strong role in the solution specifically for small values of isolation and the two solvers have different methods of meshing that
could change the overall magnitude of the isolation. The difference at 18 GHz for HFSS was a result of using rectangular vias in place of cylinder vias; however, running that frequency point with cylinder vias showed about 10 dB higher isolation. Even though these differences are notable in the plot of Figure 4.15, the isolation improvement validates that the improvement is physical and not numerical. Obtained results are encouraging and motivate the fabrication of the connected patches RIS for prototype testing.

4.3.2 Fabrication and Measurements on a Small Flat Ground Plane

The connected patches RIS was fabricated with the dimensions shown in Table 4.2. The RIS arrays were designed on several small boards (19 cm × 16.5 cm) shown in Figure 4.16 in order to compare the different RIS array sizes such that they can be tested for increasing width (41 × 21 → 41 × 41 → 41 × 81) and increasing length (21 × 41 → 41 × 41 → 81 × 41). Note that the convention used here is first number refers to the number of unit cells between the antennas (length of the array) and the second number refers to the number of unit cells that span the width of the board (width of the array). This convention of array length by width is used throughout the discussion of these studies.

The probes that are seen in Figure 4.16 are simple RF launch connectors (Omni Spectra XA2066-1401-00 [112]) which are shown along with the antenna’s return loss in Figure 4.19. Even though SMA connectors are typically limited in frequency to around 20 GHz, the small pin will still radiate and act as an antenna (small loaded monopole) at higher frequencies. It appears this monopole resonates around 24 GHz and is well matched from 20-30 GHz on all ground types. To set a baseline, a metallic ground is also measured as shown in Figure 4.17.

The measurement results for the different RIS array sizes are compared in Figure 4.18 which qualitatively validate the numerical studies and observations by showing that as the array size is increased, the isolation improves. However, in the region where the antenna is well matched (20-30 GHz), it is observed that the smaller RIS arrays do not improve the isolation as expected. After some reconsideration of these results, it is determined that this is due to the excitation of strong
Figure 4.16: Fabricated RIS on different 19 cm × 16.5 cm printed circuit boards. The boards were fabricated in such a way that the RIS could be tested by (a) increasing the width of the arrays, and (b) increasing the length of the arrays.

Figure 4.17: The metallic ground plane using the backside of one of the PCB’s from Figure 4.16.
Figure 4.18: The measured isolation is compared for (a) increasing width of the RIS arrays, and (b) increasing length of the RIS arrays.

standing waves inside the substrate, particularly for the small RIS arrays. To prove this, a $0 \times 0$ array on the substrate is also measured and compared with the metallic ground in Figure 4.20.

The isolation of the grounded substrate demonstrates that there are in fact strong standing waves inside the substrate that are heavily influencing the isolation. Frequencies below 30 GHz have much lower isolation with the presence of the substrate. This demonstrates that through comparison of the RIS to the metal ground plane by itself is not necessarily fair, but should be
Figure 4.19: Omni Spectra XA2066-1401-00 [112] SMA connectors used as probes (i.e. monopoles) for testing the isolation. The return loss is plotted for three cases: the antennas in free space which is shown on the left, the metallic ground plane which is shown in Figure 4.17, and with a metallic backed substrate which is shown in Figure 4.20. The vertical line indicates 18 GHz.

Figure 4.20: Comparison between the isolation of the metal ground plane from Figure 4.17 with a $0 \times 0$ array on a metallic backed substrate demonstrating the significant standing wave in the substrate which decreases the isolation significantly at lower frequencies.

contrasted with the metallic backed substrate instead. This can be seen in the peaks of isolation for the grounded substrate case as compared to the isolation peaks for the smaller RIS arrays. Even
with the presence of the surface directly in between the antennas, it is possible that standing waves are reflecting off of the surface and around it which explains why the isolation spikes for the small RIS arrays are more frequent. If compared with the grounded substrate, the isolation is enhanced for each of the different RIS arrays, but the larger arrays still have the highest impact as expected. This also demonstrates that for practical implementation, the RIS array should occupy most of the substrate surface to prevent such standing waves from being detrimental to the isolation.

4.3.3 Practical Implementation on a Large Flat Ground Plane

To test this practical implementation of the RIS, the flat ground plane (Figure 3.30) from Chapter 3 is considered and is shown with the RIS practically implemented onto it in Figure 4.21. The array was made to be $17 \times 81$ with a 5 mm edge of substrate is preserved for physical mounting of the RIS to the ground plane; however mounting holes could be directly implemented into the RIS if the PCB was printed with such mounting holes. Even if some standing waves are formed inside the substrate, they would be relatively contained as compared to the prototype measurements presented earlier. The computed isolation is shown in Figure 4.22 demonstrating, as expected, good improvement across the entire bandwidth.

![Figure 4.21: Model of the flat ground plane with the connected patches RIS implemented in a practical way to prevent the standing wave on a substrate.](image)
Figure 4.22: Computed isolation on the flat ground plane comparing the connected patches RIS with the ground plane by itself.

Figure 4.23: Fabricated flat ground plane with holders for the RIS array.
This RIS was then implemented on the fabricated flat ground plane from Chapter 3 (see Figures 3.42 and 3.39) which is shown in Figure 4.23 alongside aluminum holders fabricated with CNC machining to make the RIS flush mountable onto the ground plane. The isolation was measured and plotted in Figure 4.24 showing strong isolation improvement and therefore validating the previous numerical results. The results also show a clear “turn on” frequency of the RIS to be located right around 18 GHz which agrees very well with the TM surface reactance of Figure 4.8. In summary, the obtained results demonstrate that RIS’s can be designed and fabricated for broadband isolation enhancement at millimeter wave frequencies.

Figure 4.24: Measured isolation on the fabricated flat ground plane with and without the connected patches RIS.

4.4 Implementation of the RIS onto the CRS

Finally, the RIS is also implemented onto a cylinder to evaluate its suitability for use on the CRS. The same $17 \times 81$ array was “wrapped” around the cylinder used in Chapter 3 as shown in Figure 4.25. Again, both E-plane and D-plane orientations are considered for the antennas and the computed isolation is plotted in Figure 4.25. The RIS demonstrates some isolation improvement; however, it is not as consistent across the band as observed in the simulations and measurements.
Figure 4.25: The isolation for both the (a) E-plane, and (b) D-plane antenna orientations of the connected patches RIS implemented on the small cylinder ground plane to mimic the CRS. The respective models are also shown for each orientation.

There are a few factors that may contribute to this degradation in performance. The RIS is no longer flat and therefore the unit cell model is no longer a perfect representation of the surface reactance. Unfortunately, this curvature cannot be modeled in the unit cell due to the way the
periodic boundary conditions are implemented, and therefore it is hard to understand its impact on the surface reactance. We do know that the tangential electric field must be zero along the transverse strips, but now that curvature has been added, this electric field may behave differently at points just off of the cylinder which will change the surface reactance accordingly. Additionally, the tangential magnetic field is also no longer perfectly parallel with the transverse direction of the surface which will also have an impact on the surface reactance. For the D-plane results, there is an additional factor that the azimuth angle will have on the surface reactance due to the unit cell’s asymmetries. This can be modeled using the Floquet analysis by changing the direction of the incident wave. The azimuth angle, $\phi$, was varied from $0^\circ$ to $90^\circ$ as shown in Figure 4.26 demonstrating a very strong dependence on the azimuth angle for this type of a surface. For an azimuth angle of $45^\circ$ which is expected from the D-plane orientation of the antennas, the parallel resonance shifts up to about $1.15f_0$ which corresponds to around 21 GHz for the frequency range of the CRS. Moreover, the surface reactance gets “pulled” closer to zero similar to what happens when the dielectric constant of the substrate increases and therefore the surface will behave less like a high impedance surface and more like a PEC at higher frequencies.

![Diagram](image)

Figure 4.26: The dependence of the TM surface reactance on azimuth angle $\phi$ of the connected patches RIS.
This asymmetric effect is strongly due to the shorting strips between the patches, and therefore can be eliminated if the TE surface mode can be suppressed using other methods. For the case of the CRS, if metal is surrounding the antenna, then the TE surface mode should not be excited as the tangential electric field will be shorted on the metallic cylinder. Therefore, the RIS can be designed without considering the TE surface reactance for this application.

4.4.1 Hexagonal RIS Design

A nearly omnidirectional RIS has already been demonstrated in [18, 19] by using hexagonal patches instead of rectangular or square creating a honeycomb structure. The geometry of the hexagonal RIS with defined parameters considered in this thesis is shown in Figure 4.27. The period, $p$, is defined by the edge length of the hexagonal unit cell, the patch edge length, $r$, is defined by the edge of the hexagonal patch, the substrate of thickness, $t$, has dielectric constant $\varepsilon_R$, and the via is defined in terms of its diameter, $d_v$.

![Figure 4.27: Geometry of the hexagonal patches RIS including the unit cell and its parameters.](image)

To simplify the design procedure of this RIS, one can constrain parametric studies by pre-selecting the substrate: 62 mil thick Rogers 5870 RT/Duroid (dielectric constant of $\varepsilon_R = 2.33$ and loss tangent of $\tan\delta = 0.0012$) [110]. While parameters $r$, $p$, and $d_v$ were all studied parametrically, the parametric sweep considering the patch edge length, $r$, is shown in we only show Figure 4.28.
The other fixed parameters of the unit cell are: period, \( p = 0.054\lambda_0 \) and via diameter, \( d_v = 0.012\lambda_0 \) where \( \lambda_0 \) is the freespace wavelength at \( f_0 = 18 \) GHz. Again, an incidence angle of \( \theta = 80^\circ \) is used to estimate the grazing wave.

\[ \lambda_0 \]

Figure 4.28: TM and TE surface reactances of the hexagonal RIS for different values of the patch edge length \( r \).

Figure 4.29: The dependence of the TM and TE surface reactances on azimuth angle \( \phi \) of the hexagonal RIS.

To center the parallel resonance at \( f_0 \), the side of the patch is selected to have an edge length of \( 0.05\lambda_0 \). It should be noted that the TE surface mode would see a capacitive surface reactance and therefore if the TE surface mode was present, this surface should not be used. Here, we assume the metallic cylinder will eliminate the TE surface mode and thus a capacitive surface reactance is acceptable. Looking closer into the geometry of the hexagonal patches, this RIS should be almost
independent of azimuth angle, \( \phi \) because the structure itself is six-way symmetric. Because of the symmetry of a hexagon, azimuth angle \( \phi = 60^\circ \) should see the same surface reactance as \( \phi = 0^\circ \); therefore, the maximum difference in surface reactance should be expected when comparing \( \phi = 0^\circ \) and \( \phi = 30^\circ \) which is shown in Figure 4.29.

These results demonstrate that the hexagonal RIS is almost completely independent of azimuth angle which is ideal for dual polarization, D-plane oriented antennas.

This hexagonal RIS was implemented on the small ground plane used as a test bench throughout this thesis and the computed isolation is shown in Figure 4.30. For comparison with the connected patches RIS, it is also included in the plot. The isolation demonstrates that not only does the hexagonal RIS provides omnidirectionality, but the isolation improvement bandwidth is improved to almost 3.5:1.

Figure 4.30: Isolation on the small ground plane comparing the hexagonal RIS (left) to the connected patches RIS (right) demonstrating an enhancement in isolation improvement bandwidth.

### 4.4.2 Hexagonal RIS on a Flat Ground Plane

This broadband, omnidirectional isolation improvement performance is desired for implementation on the cylinder of the CRS and therefore the hexagonal RIS is further studied. First, the hexagonal RIS is added to the flat ground plane shown in Figure 4.31 along with the isolation performance. The RIS is implemented as a \( 20 \times 82 \) array with a 5 mm substrate edge for mounting.
4.4.3 Hexagonal RIS on a Cylinder Ground Plane

The isolation in Figure 4.31 demonstrates the expected behavior over the bandwidth. The additional improvement compared to the connected patches RIS may be due to its larger size (smaller unit cell but same physical dimensions on the cylinder). The results show excellent promise for integration with the CRS and the hexagonal RIS is implemented on a small cylinder in a similar fashion as the connected patches RIS in section 4.4. The obtained results are plotted in Figure 4.32. For the E-plane orientation, the hexagonal RIS on the small cylinder appears to improve the lower frequency performance significantly while preserving its broadband features. For the D-plane orientation, the performance at lower frequency is again improved significantly; however, at higher frequencies, the RIS has a few points where it is no longer helping. These points around 35 and 44 GHz are peaks in the isolation of the “Cylinder Only” case which demonstrate that this is due to a standing wave on the small cylinder. This standing wave will more than likely change when implemented onto the full cylinder as demonstrated in the measurements of Chapter 3. Moreover, since these points are peaks in isolation, they should not be creating problems for the system of the CRS.

The above discussed results demonstrate that the hexagonal RIS may be used on a CRS to enhance the isolation over a broad bandwidth. The final section of this chapter validate these
Figure 4.32: The isolation for both the (a) E-plane and (b) D-plane antenna orientations of the hexagonal RIS implemented on the small cylinder ground plane to mimic the CRS. The respective models are also shown for each orientation and the connected patches RIS is also plotted for comparison.

studies through fabrication and measurements of the hexagonal RIS on the CRS. The hexagonal RIS was not fabricated for use on the flat ground plane as the numerical studies demonstrated
the omnidirectional performance of the hexagonal RIS over the connected patches RIS was more ideal for the application of the CRS. Additionally, the flat ground plane measurements with the connected patches RIS validated the numerical studies and the reactive impedance surface theory with enough confidence to move forward with fabrication of the hexagonal RIS.

4.4.4 Hexagonal RIS Fabrication and Assembly

The hexagonal RIS is fabricated with the dimensions specified in Figure 4.27 and the fabricated surface is shown in Figure 4.33. Three of these sections measuring approximately 11 cm × 15 cm were fabricated on the same panel and then shaped to the outer diameter of the CRS.

Figure 4.33: Single section of the fabricated hexagonal RIS. Three of these sections measuring approximately 11 cm × 15 cm were fabricated and were shaped to the outer diameter of the CRS as shown in the bottom left figure.
Figure 4.34: Assembly process of the hexagonal RIS on the CRS for prototype testing. Holes were drilled and tapped in the CRS that lined up with the edges of the RIS. Then the RIS was mounted using 3D printed plastic “pucks” to pull the surface down tightly. At some points along the cylinder, small gaps were seen between the ground plane of the RIS and the cylinder which could potentially create a parallel plate waveguide and therefore copper tape was applied to the two sides of the RIS to seal this gap.
In a final system, the CRS would be shaved to make the RIS flush mountable, but for proof of concept testing, mounting the RIS to the outer diameter of the CRS would be sufficient validation. The assembly of the RIS onto the CRS is shown in Figure 4.34. First, holes were drilled and tapped into the CRS to line up with the three sections of the fabricated hexagonal RIS. Next, twelve 3D printed “pucks” were used to tighten the RIS down to the cylinder. Finally, it was noticed there was still a small gap between the ground plane of the PCB and the cylinder itself which could potentially create a parallel plate waveguide between the two grounds. Copper tape is therefore used on both sides of the RIS to seal this gap and effectively force the TM surface mode to propagate on the RIS. In a final system, the RIS would be designed to be mounted flush with the cylinder without a need for copper tape or the plastic pucks.

4.4.5 Hexagonal RIS Measurements

The CRS with and without the RIS was then measured outdoors using the measurement setup shown in Figure 4.35. The frequency sweep was from 15-40 GHz using 1601 measurement
points. The intermediate frequency bandwidth (IFBW) was set to 10 Hz to ensure a low noise floor while the RF power was set to 0 dBm. The measurements presented in Figure 4.36 demonstrate a clear improvement of isolation over the entire measured bandwidth. The measured isolation qualitatively validate the numerical results of Figure 4.32b including the minimum improvement around 35 GHz which implies the standing wave observation in the model is physical. Additionally, the implementation of the hexagonal RIS on the CRS achieves the isolation goal of 80 dB with a 3 dB margin at the minimum isolation point (around 20 GHz).

![Figure 4.36: Measured isolation of the CRS with and without the hexagonal RIS demonstrating a significant improvement across the measured bandwidth.](image)

4.5 Summary

This chapter expanded upon the work in Chapter 3, by capacitively loading corrugations to improve the bandwidth performance. These loaded corrugations can be broken up to make the
impedance surface a double periodic structure and therefore creating mushroom type structures. These types of RIS’s are typically implemented using PCB’s and the unit cell can be designed taking the substrate into consideration as was demonstrated for both the connected patches and the hexagonal patches RIS’s. The connected patches RIS is designed using soft surface principles that prevents both TE and TM surface modes from propagating, and demonstrate broadband isolation improvement for E-plane oriented antennas. To accommodate for the D-plane of the CRS, the hexagonal patches RIS was designed to maintain broadband behavior with omnidirectional performance. Both RIS’s are practical for use at millimeter waves and more importantly for wide band repeater systems such as the CRS.
5.1 Thesis Summary

This thesis discussed several techniques for isolation enhancement for RF repeater systems including antenna separation, antenna redesign, engineering losses, and most specifically and most thoroughly impedance surfaces. Impedance surfaces were studied in depth throughout this thesis as they provide the most potential for isolation enhancement without requiring modification of the antennas. They also allow for high transmitting power and are environmentally more suitable than absorbers. Moreover, minimal modification to the repeater platform is needed as the impedance surface would be on the outer side of the system and conformal to the repeater’s surface.

Surface impedance is defined as the ratio of tangential electric field to the orthogonal tangential magnetic field on that surface. Isolation can be enhanced by suppressing the surface modes propagating between the antennas which can be done by creating a capacitive surface (suppresses the TM surface mode), an inductive surface (suppresses the TE surface mode), or an electromagnetic soft surface (suppresses both TE and TM surface modes). These capacitive/inductive/soft surfaces can be designed in terms of their surface impedance and practically realized using different types of reactive impedance surfaces (RIS’s). In general, there are three types of RIS structures that are found throughout literature: corrugations, pin bed, and mushroom/patch type. The expressions for the TE and TM surface impedances for PEC uniform depth corrugations based on the Fresnel coefficients are derived. These demonstrate an octave bandwidth limitation which is fundamental to the geometry of corrugations. Similar bandwidth limitations can be observed for
the pin bed, but these findings are not explicitly presented in this thesis due to their similarity to corrugations. One can consult [20,22,27,28] for further reference on the pin bed as a RIS. To improve the bandwidth of corrugations, an analytical approach from Tereshin’s work [25,26] is applied that results in tapered depth corrugations. Alternatively, metal top-loading the corrugations can increase the capacitive surface impedance bandwidth to improve isolation over wider bandwidths than traditional corrugations. This capacitive loading can easily be applied to the pin bed; which is how the mushroom type RIS’s can be designed to be broadband. The fabrication thereof is practical with a printed circuit board (PCB).

Most studies described in this thesis utilize a simple, numerical example of small ground plane with two open ended rectangular waveguides as discussed in Chapter 3 (Figure 3.8). Because all the investigated RIS’s use this example, the isolation is shown for all of them in Figure 5.1 as a fair comparison thereof.

![Figure 5.1: A comparison of computed isolation of all different impedance surfaces analyzed in this thesis when placed on a small ground plane.](image)
In general, it is seen that the different surfaces can provide various advantages such as:

- Quarter wavelength corrugations provide the highest isolation enhancement around the design frequency,

- For applications that are already embedded in a substrate or dielectric, the mushroom RIS’s present a strong advantage as they can be easily integrated into PCB’s, not to mention that these structures can achieve wider bandwidths than traditional corrugations,

- The tapered depth corrugations can be used for broadband applications with consistent isolation enhancement performance over large frequency ranges.

All of these impedance surfaces are demonstrated to work for millimeter wave (mmw) applications with available practical fabrication techniques for both planar and cylindrical ground planes. The measured results are compared in Figure 5.2, demonstrating qualitative agreement with the small ground plate summary.

![Figure 5.2: Measured isolation of the different fabricated impedance surfaces used for isolation improvement for a (a) flat ground plane in E-plane orientation and (b) 6” diameter cylindrical ground plane in D-plane orientation.](image)

The table with the summary of these results given at the end of Chapter 1 is reproduced here as Table 5.1 for easy reference. In general, the connected patches RIS has the highest isolation
enhancement for the application's wide bandwidth. The hexagonal patches RIS can be made conformal to the platform such as with the CRS while still maintaining the desired bandwidth. Quarter-wavelength corrugations perform very well at the design frequency and can be used as a high impedance surface over smaller bandwidths. The tapered depth corrugations have consistent isolation improvement performance for both flat and cylinder ground planes and therefore can be used for multiple broadband applications. Moreover, both the tapered depth and traditional corrugations provide a high power handling solution that could be a potential issue with the PCB based RIS’s. In general, it is observed that the different impedance surfaces demonstrate better isolation improvement on a flat ground plane in E-plane orientation than on the cylinder in D-plane orientation; however, all the presented solutions may still be practically implemented and meet the specifications of the CRS application. Note that the tapered depth corrugations were merely a single implementation of Tereshin’s Surface Impedance Equation, (3.29), but this kind of tapered impedance surface could be implemented using alternative impedance surfaces such as a pin bed or the mushroom structures of Chapter 4. These results improve understanding of how these impedance surfaces perform on cylinder ground planes as compared to planar surfaces. The results from Figure 5.1 and Table 5.1 again demonstrate very good qualitative agreement in terms of trend and excellent agreement with the theory presented throughout this thesis.

It is demonstrated throughout this thesis that reactive impedance surfaces can be designed and applied in mmw applications using the developed theory from literature. These techniques and theory are not novel, nor does this thesis claim to be presenting such theory as original and new to the community; however, this thesis presents this work as a practical “guidebook” or “handbook” for design and implementation of impedance surfaces specifically for isolation enhancement, including the tangible example of mmw repeater systems. Moreover, the behavior and performance of some RIS’s that are by many referred to as metasurfaces, is traced back to the concepts first discussed in the 1950’s.
Table 5.1: Summary of results presented in this thesis for different impedance surfaces on flat and cylinder ground planes. All results are measured unless denoted with an * in which case only numerical results are shown. The “Band” used for determining Maximum, Average, and Minimum Isolation Improvement is the 17.5-45 GHz range of the CRS while the Positive Isolation Improvement Bandwidth considers all measured/simulated frequencies including those that fall below 17.5 GHz ans/or above 45 GHz.

<table>
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<tr>
<th>Impedance Surface Type</th>
<th>Maximum Isolation Improvement in the Band [dB]</th>
<th>Average Isolation Improvement in the Band [dB]</th>
<th>Minimum Isolation Improvement in the Band [dB]</th>
<th>Positive Isolation Improvement Bandwidth $[f_{max}:f_{min}]$</th>
<th>Impedance Surface Maximum Thickness [mm]</th>
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<tr>
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<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Traditional Corrugations</td>
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<td>12.2</td>
<td>&lt;0</td>
<td>2.2:1</td>
<td>4.1</td>
</tr>
<tr>
<td>Tapered Depth Corrugations</td>
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<td>13.9</td>
<td>10.3</td>
<td>&gt;2.7:1</td>
<td>4.1</td>
</tr>
<tr>
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<td>15.4</td>
<td>8.5</td>
<td>&gt;2.4:1</td>
<td>1.57</td>
</tr>
<tr>
<td>Hexagonal Patches RIS*</td>
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<td>17.1</td>
<td>7.7</td>
<td>3:1</td>
<td>1.57</td>
</tr>
<tr>
<td>Cylinder Ground Plane with Antennas Oriented in D-plane</td>
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<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
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<td>&lt;0</td>
<td>2.1:1</td>
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<td>&gt;2.6:1</td>
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</tr>
</tbody>
</table>
5.2 Some Lessons Learned, Tips, and Tricks

Throughout the studies presented in this thesis, there were many lessons learned, some of which are summarized here to assist future designers in similar studies of reactive impedance surfaces.

5.2.1 Outdoor Versus Indoor Measurements

All of the measurements are performed outdoors due to the significant impact of reflections from the ceiling (even inside an anechoic chamber). This is an important lesson learned from many early experiments were invalidated after these impacts became better understood. Appendix A dives into the details concerning time domain analysis and how to perform time gating on measurements performed indoors; however, this thesis emphasizes that there may be a difference in measurements performed indoors as compared to outdoors when high gain antennas are used to measure low levels of isolation. In order to be able to apply time gating to measurements, one must consider two important things, the first being the sampling time step, \( t_s \) in terms of the measurement bandwidth \( BW = f_{\text{stop}} - f_{\text{start}} \):

\[
t_s = \frac{1}{BW} \tag{5.1}
\]

where \( f_{\text{start}} \) and \( f_{\text{stop}} \) are the respective first and last frequency of the measurement sweep. This time step determines the time axis in the Inverse Fourier Transform. Secondly, it is important to know where the reflections will come from. One must determine the path length from the antenna to the scatterer (e.g. ceiling) and back to the other antenna to determine how many sample points are needed. The more samples taken, a longer time frame will be observable in time domain. If there is an inadequate number of sample points, then the time domain analysis will result in aliasing where the data wrap around the time axis and are not physical.
5.2.2 “Stuffing” the Cylinder’s Interior

In practice, the CRS contains internal electronic components that occupy the space between the TX and RX antennas. In measurement setup, specific details as to what goes inside the CRS was unavailable, moreover the research was specifically focused on improving the isolation and not designing the full repeater system itself. The antennas and waveguide components that are used for the tests were machined in split block machining which may produce RF leakage into the cylinder.

Figure 5.3: The process of “stuffing” the cylinder with 3.2 mm aluminum and foam absorber pucks to prevent internal RF leakages from tainting the isolation measurements.
and therefore taint the obtained isolation. To mitigate this, the cylinder’s interior is “stuffed” with foam absorber and a metal puck as demonstrated in Figure 5.3.

![Figure 5.4: Isolation of the CRS with the tapered depth corrugations with and without the “stuffing” demonstrating a strong impact in the measurements from RF leakages from the waveguide components.](image)

Figure 5.4: Isolation of the CRS with the tapered depth corrugations with and without the “stuffing” demonstrating a strong impact in the measurements from RF leakages from the waveguide components.

![Figure 5.5: Coupling in time domain of the stuffed cylinder (with tapered depth corrugations) as compared to the empty cylinder demonstrating the ringing that is due to internal RF leakage of the waveguide components inside the cylinder.](image)

Figure 5.5: Coupling in time domain of the stuffed cylinder (with tapered depth corrugations) as compared to the empty cylinder demonstrating the ringing that is due to internal RF leakage of the waveguide components inside the cylinder.
To demonstrate the impact, Figure 5.4 shows the difference in isolation for the CRS with Tapered Depth Corrugations with and without the “stuffing.” The impact seen in the isolation is significant, especially at higher frequencies. This demonstrates that leakage in components may have strong impact on the isolation and should be considered when performing such measurements. All of the measured results on the cylinder presented throughout this thesis rely on this approach to ensure the isolation is only external (i.e. due to the surface current outside of the cylinder). This leakage can be further characterized with the time domain analysis (see Appendix A) as shown in Figure 5.5, where the magnitude of $S_{21}$ is plotted over the time axis to better understand where the energy is going. The time domain data show that the empty cylinder produces ringing that is above the noise floor, whereas the stuffed cylinder has a single spike at the expected time delay between the antennas and then decays back into the noise floor. This ringing in the empty cylinder case is indicative of energy flowing inside the cylinder reflecting back and forth and leaking back into the waveguide components. Suppressing the internal leakage is beyond the scope of this thesis as the focus is on designing a RIS to suppress surface currents on the outside of the repeater system; however, these results are useful for understanding isolation measurements.

5.2.3 Epoxying Components

If the leakages cannot be suppressed using pucks like what is shown in the previous section, applying a few coats of metallic epoxy over all the waveguide components should be able to help with isolation measurements in preventing the leakage from tainting the data. After a single coat of metallic epoxy is applied to the components, it is noticed the isolation of the empty cylinder matched much better with the stuffed cylinder. There is still some rippling in the time domain analysis which may be suppressed even more if additional coats of epoxy are applied to the waveguide components.

On the topic of epoxy, it is observed that metallic epoxy should be used to mount the PCB RIS’s to other metallic surfaces to create a better contact between the ground plane of the RIS and the measurement ground plane. This helps to electromagnetically seal the air gaps that may form between them. To demonstrate this, the connected patches RIS mounted on the flat ground plane
from Figure 4.23 is considered. The small sections of the RIS are fixed into aluminum “sockets” as shown in Figure 5.6. While four screws are used to clamp the RIS to the aluminum, it is observed that if metallic epoxy was applied between the ground plane of the RIS and the aluminum socket, there is a difference in isolation as shown in Figure 5.7. While the impact is small and only appears to be at some frequencies, the measurements with the epoxy are more reliable since there is electrical contact between the ground plane of the RIS and the aluminum ground plane used for the measurements.

Figure 5.6: The connected patches RIS mounted inside the aluminum sockets for isolation measurements on the flat ground plane.
5.2.4 Modeling

Numerical modeling of bi-static repeater systems such as the CRS can quickly become computationally extensive due to the electrical size of the models and fine details of the impedance surfaces. This is eluded to in Chapter 3 (see Table 3.1), and here a few tips are shared for combating the electrical size of such problems.

Understanding the Coupling Paths

As demonstrated in Chapter 3 (Table 3.1), most of the surface currents that contribute to the isolation come from the direct path between the antennas. This means that sometimes the effect of the ground plane far away from the apertures may be ignored and the model can be made smaller. To understand which paths may be ignored, numerical studies like the one presented in Chapter 3, can be performed and observing the surface current density on the ground plane may be used for decision making.

Modeling of Curved Surfaces

Care should be taken with the numerical tools for modeling curved surfaces as often, these tools will approximate curved structures with polyhedrons which can have impact on the numerical
results. In Ansys Electronics Desktop (formerly HFSS) [92], the “initial mesh settings” dialog box is shown in Figure 5.8a with the default settings. For large curved objects, it is highly recommended to check the box labeled “Apply curvilinear elements to all curved surfaces” (or at least apply curvilinear elements to the specific object) to get a more accurate meshing on that curved structure. Additionally, the “Normal Deviation” default setting is $22.5^\circ$, which we would recommend making no more than $10^\circ$ for better mesh resolution on cylindrical surfaces. For the most accurate mesh, both options should be used; however, numerical studies can be done to see how the tolerances of the mesh impact the results. This will more than likely lead to a trade off study between the quality of the results and computational resources required. These mesh settings are also seen to improve the convergence for the unit cell/Floquet analysis for the cylindrical vias. In FEKO [111],

![Figure 5.8: (a) Ansys Electronics Desktop (formerly HFSS) [92] default Initial Mesh Settings dialog box. (b) FEKO [111] default Advanced Mesh Settings dialog box.](image)

when curvilinear mesh elements are set to auto in the default settings (see Figure 5.8b), it is found that these curved surface meshing did not have much of an issue with the solution for the cylinder
studies in Chapter 4 (see Figure 4.11). However, what is important in FEKO is to use double precision when dealing with small levels of coupling like herein. This feature can be found in the “Solver Setting” dialog box where the default is single precision.

**Boundary Conditions in FEM Solvers**

The use of hybrid solvers can be helpful in large models. For instance, the radiation boundary condition in Ansys Electronics Desktop [92] requires at least a quarter wavelength of air box between the model and the boundary. The Finite Element Boundary Integral (FEBI) hybrid boundary condition [93] can allow for this air box to be made smaller which will require less tetrahedron for meshing and therefore a quicker computation time. However, we recommend that an initial numerical study be performed to determine how far away the FEBI boundary should be from the model before it starts to significantly impact the isolation results. As a general rule of thumb, no less than one eighth of the lowest frequency wavelength should be used for this distance; however, this can be model specific and that is why the studies should be performed. For the small ground plane in Figure 5.1, a larger airbox is used because the model is small and ran relatively quickly. For the large flat ground plane, an airbox that is one sixth of the lowest frequency wavelength is used. For the small cylinder, a quarter wavelength at the lowest frequency airbox is used. Again, these numerical studies should be performed to understand how the radiation boundary condition may impact the numerical solutions.

**Multilevel Fast Multipole Method**

In FEKO [111], the multilevel fast multipole method (MLFMM) can be used for solving computationally larger problems; the “use MLFMM” check box can be found under the “Solver Settings” dialog box. It will usually yield a quicker solution with less RAM requirements than the true MoM solver; however, it is recommended to validate your results with true MoM. This validation can usually be performed on the ground plane only model.

Additionally with both of these solvers, it may often be advantageous to perform some initial study using Hertzian sources or simple wire antennas which will release some of the computational resources required for the complex antennas with greater geometrical and material detail. This
is done with the initial studies of the connected patches RIS on the cylinder in Chapter 4 (see Figure 4.11) where monopoles are used instead of implementing the horn antennas into that model.

5.3 Conclusions

This thesis demonstrated that reactive impedance surfaces can be engineered to enhance isolation over broad bandwidths on cylindrical structure mmw repeaters. The application of the cylinder repeater system presented an issue of isolation that needed solutions that do not require modification to the repeater’s antennas, internal system, environment, or power capabilities. After researching different techniques, conformal reactive impedance surfaces that are able to meet the specification of the cylinder repeater system are investigated. Theoretical understanding is demonstrated and practical implementation is carried out to achieve at least 80 dB of isolation over the required bandwidth and separation between the antennas. Additionally, less known analytical methods for antenna decoupling from the 1960’s are studied and applied to create a broadband solution with significant potential for actual use. Fabrication and measurements of such surfaces requires great care and attention and techniques for accurately measuring the isolation are presented. This thesis may assist designers to understand some of the fundamentals of different reactive impedance surfaces, their theory of operation, design strategies, fabrication, testing, and practical implementation on realistic RF, microwave, and mmw repeater platforms.
Bibliography


[112] XMA Omni Sectra Inc. XA2066-1401-00. Part no longer supported by XMA.


Appendix A

Time Domain Analysis and Time Gating

Introduction and Hypothesis

Time gating is a technique for filtering out irrelevant (or undesired) data from measurements or other results in the time domain. In other words, a time gate is applied to the data and anything that falls outside of that gate is ignored. This becomes important when measuring high levels of isolation inside an anechoic chamber. The formed hypothesis is as stated: Reflections from the chamber ceiling significantly contribute to the coupling compared to the direct path or true coupling, particularly when the antenna directivity is high. This hypothesis is depicted visually in Figure A.1.

A simple analytical test can is used first to characterize how the reflections from the ceiling may impact the isolation. We use Friis Transmission Equation (1.1) where the reflection coefficient of the ceiling, $\rho_{\text{ceiling}}$, must be considered as well as the fact that the distance, $R$, applies to the distance from the cylinder to the ceiling and back again (roughly double the height of the ceiling). By converting all of these quantities to dB, Friis Transmission Equation now becomes:

$$ P_R[\text{dBW}] = P_T[\text{dBW}] + G_T[\text{dB}] + G_R[\text{dB}] + 20 \log \left( \frac{\lambda}{4\pi R} \right) + 20 \log_{10} |\rho_{\text{ceiling}}| $$

(A.1)

It would be natural to convert (A.1) into an expression for the isolation by subtracting $P_T$ from both sides and inverting the sign of both sides which then yields:

$$ Isolation[\text{dB}] = P_T[\text{dBW}] - P_R[\text{dBW}] = - \left( G_T[\text{dB}] + G_R[\text{dB}] + 20 \log \left( \frac{\lambda}{4\pi R} \right) + 20 \log_{10} |\rho_{\text{ceiling}}| \right) $$

(A.2)
Figure A.1: Visual depiction of the time gating hypothesis. The figure shows the CRS on the floor, the ceiling, and the propagation of different waves that result in coupling.

Figure A.2: Measured gain of the antenna plugged into (A.2) with the assumptions of $R = 4.86$ m and -30 & -40 dB for $\rho_{ceiling}$ compared with the measured isolation inside the anechoic chamber.

To find the isolation, a few simple assumptions can be substituted into (A.2). First, the gain of the TX and RX antennas ($G_T$ and $G_R$ respectively) can be substituted using the measured boresight gain. The chamber ceiling is roughly 2.4 m above the cylinder and therefore, $R$ should
be 4.8 m because the wave travels 2.4 m there and back. The reflection coefficient of the chamber ceiling is assumed to be somewhere between -30 and -40 dB [113]. Using these assumptions, the isolation is computed and compared in Figure A.2 with the measured isolation of the CRS inside the chamber.

As seen, the reflections from the absorber on the ceiling can be comparable to or even dominate the measured isolation. Unless the expected isolation is less than about 70 dB in this frequency range, the reflections from the ceiling are on the same order of magnitude of the system isolation even with good absorber (40 dB). Moreover the measurements shown are for the metallic cylinder only. After the addition of corrugations or RIS, the isolation is expected to increase by at least 15 dB, therefore the actual results will be buried in these reflections. To demonstrate this, the measured isolation of the CRS with and without the hexagonal RIS is shown in Figure A.3. The plot shows that the results are nearly on top of each other which contradicts the findings presented at the end of Chapter 4 (see Figure 4.36). The conclusion is that these measurements are tainted by the ceiling reflections instead of measuring the true isolation on the cylinder.

**Time Domain Procedure and Analysis**

The measurement data can be viewed in the time domain (TD) as compared to the frequency domain (FD) by using an Inverse Discrete Fourier Transform (IDFT). Essentially, the IDFT is a mathematical operator on a discrete FD dataset to represent the same measurable data in TD. The IDFT is also the inverse operator of the Discrete Fourier Transform (DFT) such that given a TD signal, \( y(t) \), then the following properties of the DFT and IDFT must hold true:

\[
Y(f) = \text{DFT}[y(t)] \\
y(t) = \text{IDFT}[Y(f)] \tag{A.3}
\]

This tool has been implemented in most coding languages, one of these being MatLab [104]; which is used in this thesis for data processing. MatLab has implemented the DFT and IDFT using a Fast Fourier Transform (FFT) algorithm [104]. This FFT (and it’s inverse: IFFT) are used to analyze the measured isolation in the TD. During the IFFT, it is important to know how to map
the frequency axis to the time axis which is dependent on the frequency step taken during the measurement. The IFFT will only compute the TD data from $t = 0$ to $T_{\text{end}}$ with:

$$T_{\text{end}} = \frac{N - 1}{BW}$$  \hspace{1cm} (A.4)

where $N$ is the number of frequency points and $BW = f_{\text{stop}} - f_{\text{start}}$ is the bandwidth of the measurement: $f_{\text{start}}$ and $f_{\text{stop}}$ being the respective first and last frequency points. As an example, the measurements shown in Figure A.3 were taken from 15-40 GHz with 501 frequency points, therefore $f_{\text{start}} = 15 \text{ GHz}$, $f_{\text{stop}} = 40 \text{ GHz}$, $BW = f_{\text{stop}} - f_{\text{start}} = 40 - 15 = 25 \text{ GHz}$, $N = 501$, thus $T_{\text{end}} = \frac{500}{25 \text{ GHz}} = 20 \text{ ns}$. Care should be taken to avoid a signal with significant energy close to $T_{\text{end}}$ or later otherwise it can wrap around the time axis and appear at $t \approx 0$, and aliasing the measured data. This stresses the importance of selecting the adequate frequency step ($N$ for a fixed bandwidth), for isolation measurements inside an anechoic chamber. For free space propagation, the time delay, $\tau_0$ of a signal traveling a distance, $R$ is:

$$\tau_0 = \frac{R}{c_0}$$  \hspace{1cm} (A.5)
where \( c_0 \) is the speed of light in free space. Considering the anechoic chamber used in these experiments, which has a ceiling height of 2.4 m, and therefore \( R = 4.8 \) m, then the time delay should be around \( \tau_0 = 16 \) ns. By setting \( T_{end} \geq \tau_0 = 16 \) ns and solving (A.4) for \( N \) tells us that \( N \geq 406 \) measurement points, when a 15-40 GHz frequency range is used.

Performing an IFFT on the frequency data of the cylinder only case from Figure A.3 and the magnitude of the isolation in TD is shown in Figure A.4. There are two significant spikes in the TD, one occurring at 2.5 ns and the second around 18 ns. Because there is not an adequate number of points used in this measurement, the tail of the second spike wraps around the time axis and appears to have been shifted to 0 ns where the tail continues to decay. This signal that has wrapped to 0 ns is not physical and is an aliasing of the second signal due to the property of the FFT. The spike that begins around 18 ns is relatively close to the 16 ns time delay expected from the chamber ceiling. When considering the first spike in the data, it begins around 2.5 ns which compared with the expected time delay for the antenna separation (2.4 m \( \rightarrow \tau_0 = 0.7 \) ns) is also shifted by about 2 ns. Because the free space speed of light is roughly 30 cm/ns, a 2 ns shift is not insignificant and needs to be accounted for. It turns out this 2 ns shift is due to the calibration plane of the measurement setup. For these measurements, a standard 2.4 mm coaxial SOLT calibration is used to calibrate the network analyzer which means the calibration plane does not take into account the coax-to-waveguide adapters, orthomode transducers (OMT’s) with phase matched lines, and the path length of the horn antenna itself. It turns out that for the CRS, the extra time the signal takes to travel through these components for both antennas combined is roughly 1.75 ns which is shown later in this appendix. We refer to this extra added time as the “Aperture Delay” because it is the combined time shift it takes for the signal to reach both apertures.

Now that the TD plot has been interpreted, signal processing can be performed on the TD data to separate the physical (true) from the artificial (reflected off the environment) isolation. The signal processing performed here is referred to as “Time Gating” because essentially the TD data is filtered or “gated” and is then transformed back into FD. This processing is gating out signals that take a relatively long time to arrive at the receiver. For these studies, a simple rectangular
window function \((w(t))\) is used as the gate meaning that in the time window of interest the signal is multiplied by 1 while outside of this window, the signal is multiplied by 0. This is visually demonstrated in Figure A.5 where the window is set to gate the time window between \(t = 1.75\) ns and \(t = 15\) ns.

After this gating is performed, an FFT of the time gated data shows the (expected) true isolation which is been plotted along with the raw (ungated) isolation in Figure A.6. There is clearly a great improvement in isolation due to the RIS seen in the gated data, but obscured in
Figure A.6: Isolation of the CRS before and after time gating showing how the reflections from the chamber ceiling do in fact have an impact on the isolation.

The original measurements. Comparing the curves of Figure A.6 shows that there is a form of smoothing happening to the curve, but also at higher frequencies, the magnitude is significantly impacted. Particularly around 37 GHz, where the true isolation is greater than 90 dB, but the measurements showed about 78 dB of isolation. This validates the results presented in Figure A.2 showing that the reflections from the absorber will taint the measurements, even if the reflection coefficient of the absorber is on the order of -40 dB.

This same procedure (including the same time window) is applied to the CRS with the hexagonal RIS and the results are compared in Figure A.7. Comparing the plot on the left to the plot on the right of Figure A.7, there is good agreement for the cylinder only case demonstrating that this time gating is working; however, the cylinder with the RIS shows the right order of magnitude, but the trend does not seem to agree very well. To debug this issue, the measured data is viewed in the TD:

Figure A.8 shows that the tail of the chamber reflections wrapping around the time axis can still contaminate the measurements, even after time gating. Notice that the magnitude of the signal with the RIS is much smaller compared to the cylinder only as well as the reflections from
Figure A.7: The isolation of the CRS with and without the RIS. The measured results from outside are shown on left (note that this measurement used 801 samples), the raw measured data of the isolation from inside the chamber before time gating (501 samples) shown in the center, and the isolation data after the time gating was applied shown on the right.

Figure A.8: The time domain data of the measured isolation of the CRS with the RIS both before (left) and after (right) the time gate (same window function from Figure A.5) has been applied. These measurements were taken with 501 frequency samples and therefore the reflections from the ceiling are wrapping around the time axis and contaminating the isolation measurement.

the ceiling. This means that the tail end of the ceiling reflections must be taken into account before the measurement setup. The experiments are performed again with 751 points which results in a $T_{end} = 30$ ns. The TD data is shown for both the CRS with and without the RIS for these new
measurements before and after the time gating (same window function from \( t = 1.75 \text{ ns} \) to 15 \text{ ns}) in Figure A.9.

![Figure A.9: The time domain data of the measured isolation of the CRS with and without the RIS both before (left) and after (right) the time gate (same window function from Figure A.5) has been applied. These measurements were taken with 751 frequency samples and therefore the reflections from the ceiling decay into the noise floor before wrapping around the time axis and do not interfere with the true isolation.](image)

These results demonstrate even more clearly that the RIS has a significant impact on the true isolation (first spike), but very little impact on the reflections from the chamber (second spike). After the time gating is performed, the data is brought back into FD and the isolation is compared in Figure A.10. These results show a much better agreement with the outdoor measurements due to the fact that the tail of the ceiling reflections has decayed into the noise floor of the measurements by the time it wraps around the time axis.

To further validate this, Figure A.11 compares the TD data of the outdoor measurements side-by-side with the indoor measurements. The plot shows that everything left of 15 \text{ ns} is nearly identical to both plots which is the expected/true isolation, while the reflections from the chamber ceiling only appear in the indoor measurements. Additionally, if comparing the right hand side plot of Figure A.11 (outdoor measurements) with the right hand side plot of Figure A.9 (indoor
measurements after time gating), there is also good agreement in the TD again validating the time
gating approach to indoor isolation measurements.

Figure A.10: The isolation of the CRS with and without the RIS. The measured results from
outside are shown on left (note that this measurement used 801 samples), the raw measured data
of the isolation from inside the chamber before time gating (751 samples) shown in the center, and
the isolation data after the time gating from Figure A.9 was applied shown on the right.

Figure A.11: The time domain of the CRS with and without the RIS comparing side-by-side the
indoor measurements (left) and the outdoor measurements (right).

The small disagreements between the outdoor and indoor measurements are contributed
to three things: PNA calibration (temperature, environment, etc.), physical measurement setup
(copper tape, cable position, etc.), and the time gate itself. The first two are fairly straightforward, the network analyzer’s calibration will take into account the cable length as well as the machine’s current status such as temperature. As the machine is moved from outdoors to indoors or vice-versa, even with re-calibration, the environment may impact the precision of the calibration, particularly when measuring signals on the order of -80 dB. The second possible source of error comes from the measurement setup. This includes the copper tape that has been applied to the RIS to seal the gap between the RIS’s ground plane and the cylinder. The copper tape is unlikely applied in the exact same way every time; moreover, the wrinkles in the copper tape can create a surface roughness that will change the way the surface mode propagates along the cylinder therefore altering the isolation. Additionally, small shifts in the cables may cause small reflections in the measurements that may impact the isolation when measuring small signals.

Thirdly, the time gate itself can create some error in the post-processed measurement data. When the time gate is applied in TD, we are multiplying two functions: let \( i(t) \) represent the TD isolation signal and the TD window function has already been defined as \( w(t) \) such that the time gated isolation function is \( i_{TG}(t) = i(t)w(t) \). These functions have Fourier Transforms \( I(f) \) and \( W(f) \) respectively; however, their product TD results in a convolution in FD: \( I_{TG}(f) = I(f) * W(f) \). Where \( * \) is the convolution operator defined as:

\[
I_{TG}(f) = I(f) * W(f) = \int I(\gamma)W(f - \gamma)d\gamma. \tag{A.6}
\]

In the case of the time gating window function \( (w(t)) \), its Fourier Transform will be sinc function \( \text{sinc}(x) = \frac{\sin(x)}{x} \) which has an infinite frequency support. This means that after the measured isolation has been convolved with this sinc function, there may be some nonphysical effects associated with the sidelobes of the sinc function that extend over all frequencies and the finite/limited support of the measured data. In other words, it is possible this convolution will not perfectly reconstruct the true isolation.

We did some initial investigation into different window functions and a few are compared in Figure A.12 which include the rectangular window used earlier, a sinc function in time domain,
Figure A.12: The three window functions overlayed on the measured data used for time gating showing the sinc function (left), the rectangular window function (center), and Gaussian pulse (right).

and a Gaussian pulse. The time gated isolation data is compared in Figure A.13 which in general shows small impact on the time gated isolation data, but still there is an impact based on which window function is chosen due to the nonphysical effects associated with the convolution. There has been significant work done found in literature to understand these effects including the use of alternative window functions [114–117]; however, the observed results gave confidence that the approach adequately approximated the true isolation for this thesis.

Figure A.13: The isolation after time gating comparing the different window functions from Figure A.12
Finding the Aperture Delay

Exactly knowing the aforementioned aperture delay could become crucial in certain circumstances where scatterers are positioned such that time gating needs to start exactly when the signal reaches the aperture. To find this aperture delay, we only need to take two measurements: one with the antenna apertures open and one with the apertures covered in aluminum tape as shown in Figure A.14.

![Open Aperture](open_aperture.png) ![Covered Aperture](covered_aperture.png)

Figure A.14: The time domain of the CRS with and without the RIS comparing side-by-side the indoor measurements (left) and the outdoor measurements (right).

The time domain analysis can be performed on the port reflection coefficients to see when the wave reflects off of the tape covering. Overlaying the time domain plots for the open and closed will allow for the aperture delay to be determined. The TD plots of the port reflection coefficients ($S_{11}$ and $S_{22}$) are shown in Figure A.15. The time at which the covered aperture and the open aperture begin to disagree is the time it takes for the signal to reflect off the aluminum tape and reflect back into the port which has been indicated with a vertical line in the plots. The aperture delay of each port corresponds to half of this time because we are measuring the reflection coefficient which means it is the time it takes for the signal to travel through the waveguide components and back. Therefore the total aperture delay will be half of the sum of the time for both ports. For this case, Port 1 had a reflection delay of 1.93 ns while for Port 2 it was 1.55 ns, therefore the total aperture delay equals $\frac{1.93 \text{ ns} + 1.55 \text{ ns}}{2} = 1.74 \text{ ns}$. 
Figure A.15: The time domain of the CRS’s port reflection coefficients comparing the open apertures with the covered apertures. Port 1 ($S_{11}$) is shown on the left while Port 2 ($S_{22}$) is shown on the right.
A.1 Code for Time Gating

The MatLab code for the time gating has been included in this appendix. This code takes a touchstone format file and performs time gating using a rectangular window function based on the user input parameters.

```matlab
1 clear all; % Clear variables
2 eps0 = 8.85e−12; mu0 = 4*pi*1e−7; c0 = 1/sqrt(eps0*mu0); % Physical Constants
3 % SETUP & INPUT PARAMETERS
4 t1 = 1.75e−9; % t1 = 1.75 ns
5 t2 = 15.0e−9; % t2 = 15.0 ns
6 File = 'path/filename.sNp'; % Enter filename here (touchstone format)
7 % Perform time domain analysis on port parameter Sij
8 port_i = 2;
9 port_j = 1;
10
11 % READ FILE
12 S = sparameters(File); % Read file
13 Sparam(:,1) = S.Parameters(port_i, port_j,:); % Pull Sij for analysis
14 f = S.Frequencies;
15 N = length(f); % Number of Points
16 % Convert to Time Domain
17 T = (N−1)/(f(end) − f(1)); % Calculate T_end
18 Y = ifftshift(Sparam);
19 Tparam = ifft(Y,N);
20 t = 0:T/(N−1):T; % Generate time axis
21 % Generate Window Function w(t) = {1 if t1 < t < t2 , 0 else}
```
for j = 1:length(Tparam)
    ti = t(j);
    if ((ti <= t2) && (ti >= t1))
        U(j) = 1;
    else
        U(j) = 0;
    end
end
U = U';
% Apply Window Function
Tparam_TG = U.*Tparam;
% Recreate S–Parameter in Frequency Domain
Y = fft(Tparam_TG,N);
Sparam_TG = fftshift(Y);

%% PLOTTING
% Plot Window Function: w(t)
figure; hold on; grid on;
set(gca, 'FontSize', 16);
xplt = t*1e9;
yplt = U;
plot(xplt, yplt);
xlabel('Time [ns]');
ylabel('w(t)');
% Plot With Window: Tparam(t), w(t)
figure; hold on; grid on;
set(gca, 'FontSize', 16);
xplt = t*1e9;
yplt = abs(Tparam);
plot(xplt, yplt);

yplt2 = U*max(yplt);
plot(xplt, yplt2);
xlabel('Time [ns]');
ylabel('|f(t)| \cdot w(t)');
legend({'f(t)', 'w(t)'});

% Plot Multiplied by Window Tparam(t).w(t)
figure; hold on; grid on;
set(gca, 'FontSize', 16);
xplt = t*1e9;
yplt = abs(Tparam_TG);
plot(xplt, yplt);
xlabel('Time [ns]');
ylabel('|f(t)| \cdot w(t)');

% Plot original Sparam with Time Gated Sparam
figure; hold on; grid on;
set(gca, 'FontSize', 16);
xplt = f*1e-9;
yplt = 20*log10(abs(Sparam));
plot(xplt, yplt);

yplt2 = 20*log10(abs(Sparam_TG));
plot(xplt, yplt2);
xlabel('Frequency [GHz]');
76  ylabel('S-Parameter [dB]');

77  legend({'Before Time Gating', 'After Time Gating'});
Appendix B

Noise Floor Measurements

Introductory Noise Floor Theory

The measurements performed throughout this thesis are of very small power levels, with the isolation being on the order of 80 dB. Measurements of such power levels need to be taken with care to ensure that the signal is not buried beneath the noise floor. If the noise floor is several orders of magnitude beneath the measured signal, then the confidence in the measurement accuracy is high. In other words, it is desired to maximize the signal to noise ratio (SNR) throughout these measurements.

The most important aspect of these noise measurements is the setup and calibration of the network analyzer. For all of the measurements performed throughout this thesis including the noise measurements presented in this appendix, an Agilent 1 E8363B PNA series network analyzer is used. From noise theory [6, 118, 119], the thermal noise power ($P_n$), also called white noise or Johnson-Nyquist noise, generated in a circuit at temperature, $T$ in Kelvin is given by:

$$P_n = k_B T B$$  \hspace{1cm} (B.1)

where $k_B = 1.3806 \cdot 10^{-23}$ J/K is the Boltzmann Constant in units of Joules/Kelvin and $B$ is the bandwidth of the signal. This bandwidth corresponds to the Intermediate Frequency Bandwidth (IFBW) which is the bandwidth of the filter at baseband during the processing of the signal inside the PNA. Because of the directly proportional relationship between $P_n$ and IFBW, a narrow IFBW should be used to push the noise floor to lower power levels. From (B.1), the spectral density of the

\footnote{formerly Agilent Technologies which is now Keysight Technologies}
noise is simply given as $k_B T$ which equates to about -174 dBm/Hz. Table B.1 gives the thermal noise floor in dBm for different values of the IFBW.

Table B.1: Thermal noise power versus IFBW

<table>
<thead>
<tr>
<th>IFBW</th>
<th>Noise Power, $P_n$ [dBm]</th>
<th>$P_n,\text{dBm} - (-174 \text{ dBm})$ [dB]</th>
<th>10 log$_{10}$ (IFBW/1 Hz) [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>1 Hz</td>
<td>-174</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>10 Hz</td>
<td>-164</td>
<td>10</td>
<td>10</td>
</tr>
<tr>
<td>20 Hz</td>
<td>-161</td>
<td>13</td>
<td>13</td>
</tr>
<tr>
<td>50 Hz</td>
<td>-157</td>
<td>17</td>
<td>17</td>
</tr>
<tr>
<td>100 Hz</td>
<td>-154</td>
<td>20</td>
<td>20</td>
</tr>
<tr>
<td>200 Hz</td>
<td>-151</td>
<td>23</td>
<td>23</td>
</tr>
<tr>
<td>500 Hz</td>
<td>-147</td>
<td>27</td>
<td>27</td>
</tr>
<tr>
<td>1 kHz</td>
<td>-144</td>
<td>30</td>
<td>30</td>
</tr>
<tr>
<td>2 kHz</td>
<td>-141</td>
<td>33</td>
<td>33</td>
</tr>
<tr>
<td>5 kHz</td>
<td>-137</td>
<td>37</td>
<td>37</td>
</tr>
<tr>
<td>10 kHz</td>
<td>-134</td>
<td>40</td>
<td>40</td>
</tr>
</tbody>
</table>

Measuring thermal noise is not trivial due to the extremely small levels of power the receiver needs to be sensitive to, particularly because the measurement machine (in this case the PNA) will have internal active (LNAs, mixers, etc.) and passive (filters, lossy cables, etc.) components that will add additional noise into the system. Each of these components has its own Noise Figure ($NF$) which is defined as the ratio of the $SNR$ at the input to the $SNR$ at the output:

$$NF \text{ [dB]} = 10 \log_{10} \left( \frac{SNR_{in}}{SNR_{out}} \right) = SNR_{in, dB} - SNR_{out, dB}. \quad (B.2)$$

The $NF$ of passive components is simply equal to their insertion loss, whereas for active components, these $SNR$ ratios need to be measured to determine the $NF$. As these components are added to the measurement, they will insert additional noise that will increase the noise from the thermal noise floor. At some point, these additional noise sources will limit the PNA’s receiver and create an internal noise floor that is higher than thermal noise. This is usually referred to as the machine’s noise floor. Measuring the machine’s noise floor is easier because the individual internal noise figures do not have to be taken into account because they are considered part of the noise floor.
Noise Floor Measurement Method, Calibration, and Procedure

To estimate the machine noise floor in the experiments in this thesis, an isolation measurement (measuring $S_{21}$ or $S_{12}$) while both ports are theoretically perfectly isolated (independently terminated) is utilized. Cable leakage may contribute to this noise floor; however, it is ignored for the purposes of these experiments. Investigating the isolation measurement data in time domain (see Appendix A) may be used to determine if the cable leakage is a significant contribution to the noise floor. During calibration of the PNA, the option to “Omit Isolation” is often checked by default because the uncorrected port isolation is usually adequate for most component measurements; however, for devices with “high insertion loss”, the isolation calibration should be included as part of the calibration process [120,121]. By opting into an isolation calibration, the PNA will perform an additional sweep during the “Load” standard of a SOLT calibration for each port to measure the coupling between the ports and then correct for this coupling [120,121]. During this isolation calibration, ALL ports should be terminated with a good impedance match [120,121]! If only one set of standards are available, use the standard on the port that is being tested and terminate the other port(s) with an alternative good impedance match.

The procedure for these noise measurements is straightforward and described in three steps:

1) Select IFBW

2) Calibrate PNA (including Isolation Calibration)

3) Terminate Both Ports and Measure

For the measurements presented here, 6 various IFBW’s are tested, while the additional PNA setup parameters are: frequency sweep from 15-40 GHz, 801 points, RF power of -12 dBm, and no averaging.
Noise Floor Measurement Results

The measured results are presented in Figure B.1. Due to the nature of the noise, it is difficult to observe what is happening when the raw data is plotted on top of each other as shown on the left side of Figure B.1. For visual aid, each of the curves were “smoothed” with MatLab’s smoothing function [104] which is shown on the right side. Clearly seen, there is a drop in isolation at 20 GHz that is independent of the IFBW. This is attributed to internal functioning of the PNA. Likely from switching components that cover different frequency ranges. More importantly, the results demonstrate that as the IFBW increases, the noise floor increases as is expected from (B.1). This change is most obvious by comparing the average value across the measured frequency range in which a linear relationship between the IFBW and the noise power is observed. Table B.2 compares the average noise as well as the figure of merit, ∆ Noise, which is the average noise for each IFBW minus (in dB) the average isolation noise for the lowest IFBW (10 Hz in this case). This ∆ Noise parameter shows extremely good agreement with (B.1) but the average noise is shifted in magnitude.
due to the noise figures of the PNA. Increasing the IFBW 10 Hz to 20 Hz (factor of 2 in IFBW),
the noise floor increases by 3 dB (factor of 2), whereas increasing it from 10 Hz to 100 Hz (factor
of 10), the noise increases by 10 dB (factor of 10) demonstrating that this method is accurately
measuring the PNA’s noise floor.

Table B.2: Average measured noise floor versus IFBW

<table>
<thead>
<tr>
<th>IFBW</th>
<th>Average Noise [dB]</th>
<th>∆ Noise [dB]</th>
<th>$10 \log_{10} (IFBW/10 \text{ Hz})$ [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>10 Hz</td>
<td>-107.1</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>20 Hz</td>
<td>-104.1</td>
<td>3.00</td>
<td>3</td>
</tr>
<tr>
<td>50 Hz</td>
<td>-100.1</td>
<td>7.00</td>
<td>7</td>
</tr>
<tr>
<td>100 Hz</td>
<td>-97.28</td>
<td>9.82</td>
<td>10</td>
</tr>
<tr>
<td>1000 Hz</td>
<td>-86.89</td>
<td>20.21</td>
<td>20</td>
</tr>
<tr>
<td>10,000 kHz</td>
<td>-76.47</td>
<td>30.63</td>
<td>30</td>
</tr>
</tbody>
</table>
## Appendix C

### Table of Symbols and Acronyms

<table>
<thead>
<tr>
<th>Symbol or Acronym</th>
<th>Name</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>AMC</td>
<td>artificial magnetic conductor</td>
<td></td>
</tr>
<tr>
<td>$\alpha$</td>
<td>attenuation constant</td>
<td>units of Nepers per meter</td>
</tr>
<tr>
<td>$\beta$</td>
<td>phase constant</td>
<td>units of radians per meter</td>
</tr>
<tr>
<td>CRS</td>
<td>Cylinder Repeater System</td>
<td>name of application used throughout this thesis</td>
</tr>
<tr>
<td>$C$</td>
<td>capacitance</td>
<td>units of Farads, capacitance implies negative reactance: $X &lt; 0$</td>
</tr>
<tr>
<td>$c_0$</td>
<td>speed of light in vacuum</td>
<td>units of meters per second, $c_0 = \frac{1}{\sqrt{\varepsilon_0 \mu_0}} \approx 3 \cdot 10^8$ m/s</td>
</tr>
<tr>
<td>$\delta_s$</td>
<td>skin depth</td>
<td>units of meters</td>
</tr>
<tr>
<td>DFT</td>
<td>Discrete Fourier Transform</td>
<td></td>
</tr>
<tr>
<td>D-plane</td>
<td>Diagonal-plane</td>
<td>antennas oriented such that the electric fields are at a 45° with the path between them</td>
</tr>
<tr>
<td>Symbol</td>
<td>Description</td>
<td>Units</td>
</tr>
<tr>
<td>--------</td>
<td>-------------</td>
<td>-------</td>
</tr>
<tr>
<td>$E$</td>
<td>electric field</td>
<td>units of Volts per meter</td>
</tr>
<tr>
<td>EBG</td>
<td>electromagnetic bandgap</td>
<td></td>
</tr>
<tr>
<td>E-plane</td>
<td>antennas oriented such that the electric fields are parallel with the path between them</td>
<td></td>
</tr>
<tr>
<td>$\epsilon_0$</td>
<td>permittivity of free space</td>
<td>units of Farads per meter, $\epsilon_0 \approx 8.85 \cdot 10^{-12}$ F/m</td>
</tr>
<tr>
<td>$\eta_0$</td>
<td>impedance of free space</td>
<td>units of Ohms, $\eta_0 = 120\pi \approx 377 \Omega$</td>
</tr>
<tr>
<td>$f$</td>
<td>frequency</td>
<td>units of Hertz</td>
</tr>
<tr>
<td>FD</td>
<td>Frequency Domain</td>
<td></td>
</tr>
<tr>
<td>FEBI</td>
<td>finite element boundary integral</td>
<td></td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier Transform</td>
<td></td>
</tr>
<tr>
<td>$G_R$</td>
<td>gain of the receiver</td>
<td>usually in units of dB</td>
</tr>
<tr>
<td>$G_T$</td>
<td>gain of the transmitter</td>
<td>usually in units of dB</td>
</tr>
<tr>
<td>$H$</td>
<td>magnetic field</td>
<td>units of Amps per meter</td>
</tr>
<tr>
<td>HIS</td>
<td>high impedance surface</td>
<td></td>
</tr>
<tr>
<td>HIT</td>
<td>high impedance threshold</td>
<td></td>
</tr>
<tr>
<td>H-plane</td>
<td>antennas oriented such that the electric fields are perpendicular with the path between them</td>
<td></td>
</tr>
<tr>
<td>IBC</td>
<td>impedance boundary condition</td>
<td></td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
<td></td>
</tr>
<tr>
<td>--------------</td>
<td>-------------</td>
<td></td>
</tr>
<tr>
<td>IDFT</td>
<td>Inverse Discrete Fourier Transform</td>
<td></td>
</tr>
<tr>
<td>IFBW</td>
<td>Intermediate Frequency Bandwidth usually in units of Hertz</td>
<td></td>
</tr>
<tr>
<td>$k$</td>
<td>Wavenumber units of inverse meters</td>
<td></td>
</tr>
<tr>
<td>$L$</td>
<td>Inductance units of Henries, inductance implies positive reactance: $X &gt; 0$</td>
<td></td>
</tr>
<tr>
<td>$\lambda$</td>
<td>Wavelength usually in units of meters</td>
<td></td>
</tr>
<tr>
<td>MLFMM</td>
<td>Multilevel fast multipole method</td>
<td></td>
</tr>
<tr>
<td>MM</td>
<td>Metamaterial</td>
<td></td>
</tr>
<tr>
<td>mmw</td>
<td>Millimeter wave</td>
<td></td>
</tr>
<tr>
<td>$\mu_0$</td>
<td>Permeability of free space units of Henries per meter, $\mu_0 = 4\pi \cdot 10^{-7}$ H/m</td>
<td></td>
</tr>
<tr>
<td>$\omega$</td>
<td>Angular frequency units of radians per meter, related to frequency through: $\omega = 2\pi f$</td>
<td></td>
</tr>
<tr>
<td>OMT</td>
<td>Orthomode transducer</td>
<td></td>
</tr>
<tr>
<td>PCB</td>
<td>Printed circuit board</td>
<td></td>
</tr>
<tr>
<td>PEC</td>
<td>Perfect electric conductor</td>
<td></td>
</tr>
<tr>
<td>$\phi$</td>
<td>Phase units of radians or degrees</td>
<td></td>
</tr>
<tr>
<td>PMC</td>
<td>Perfect magnetic conductor</td>
<td></td>
</tr>
<tr>
<td>Pol</td>
<td>Polarization</td>
<td></td>
</tr>
<tr>
<td>PPWG</td>
<td>Parallel plate waveguide</td>
<td></td>
</tr>
<tr>
<td>Symbol</td>
<td>Description</td>
<td>Unit Notes</td>
</tr>
<tr>
<td>--------</td>
<td>-------------</td>
<td>------------</td>
</tr>
<tr>
<td>$P_R$</td>
<td>received power</td>
<td>usually in units of Watts or dBm</td>
</tr>
<tr>
<td>$P_T$</td>
<td>transmitted power</td>
<td>usually in units of Watts or dBm</td>
</tr>
<tr>
<td>$R$</td>
<td>distance or radius</td>
<td>usually in units of meters</td>
</tr>
<tr>
<td>RF</td>
<td>radio frequency</td>
<td></td>
</tr>
<tr>
<td>$\rho$</td>
<td>reflection coefficient</td>
<td></td>
</tr>
<tr>
<td>RIS</td>
<td>reactive impedance surface</td>
<td></td>
</tr>
<tr>
<td>RX</td>
<td>receiver</td>
<td></td>
</tr>
<tr>
<td>$\sigma$</td>
<td>conductivity</td>
<td>units of Siemens per meter</td>
</tr>
<tr>
<td>$S_{ij}$</td>
<td>scattering parameter from port $j$ to port $i$</td>
<td></td>
</tr>
<tr>
<td>$SNR$</td>
<td>Signal to Noise Ratio</td>
<td>usually in units of dB</td>
</tr>
<tr>
<td>TD</td>
<td>Time Domain</td>
<td></td>
</tr>
<tr>
<td>TDC</td>
<td>tapered depth corrugations</td>
<td></td>
</tr>
<tr>
<td>TE</td>
<td>transverse electric</td>
<td>denotes an electromagnetic wave that propagates orthogonal to the electric field</td>
</tr>
<tr>
<td>TEM</td>
<td>transverse electromagnetic</td>
<td>denotes an electromagnetic wave that propagates orthogonal to both the electric and magnetic fields</td>
</tr>
<tr>
<td>$\theta$</td>
<td>angle of incidence</td>
<td>units of radians or degrees</td>
</tr>
<tr>
<td>TM</td>
<td>transverse magnetic</td>
<td>denotes an electromagnetic wave that propagates orthogonal to the magnetic field</td>
</tr>
<tr>
<td>TX</td>
<td>transmitter</td>
<td></td>
</tr>
<tr>
<td>$X$</td>
<td>reactance</td>
<td>units of Ohms, the imaginary part of impedance</td>
</tr>
<tr>
<td>-----</td>
<td>-----------</td>
<td>---------------------------------------------</td>
</tr>
<tr>
<td>$Z$</td>
<td>impedance</td>
<td>units of Ohms, complex number consisting of resistive (real) and reactive (imaginary) parts</td>
</tr>
</tbody>
</table>