30 GHz Switchable Waveguide Antenna:

Prototype Antenna Solution for 60 GHz Wi-Fi Protocol
802.11 ad

by

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ABSTRACT

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30 GHz Switchable Waveguide Antenna A prototype for 60 GHz Wi-Fi

The purpose of this thesis was to design an antenna for 60 GHz Wi-Fi protocol 802.11 ad. Due to high loss at this frequency, highly directional, narrow beam antennas are required. Therefore, for WLAN applications where hemispherical coverage is required, a multi-antenna-element solution is required.

In response to these requirements, a switchable waveguide structure was built by combining substrate integrated waveguide technology with controllable reflectors. Prototypes were designed and manufactured at 30 GHz to test this new technology. Prototypes included 1-way, 2-way and 4-way switchable waveguide variations. Measured results are in line with initial design targets and simulation results.

THE TEAM

Figure 1: The Team - from left to right: Peter Frank, Joel Hadden, Jim Wight, Roland Smith

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ACRONYMS

AP - Access Point
BW - Bandwidth
CMOS - Complementary Metal Oxide Semiconductor
DUT - Device Under Test EIRP - Effective Isotropically Radiated Power
EM - Electro-Magnetic
FSPL - Free Space Path Loss
FZPL - Fresnel Zone Plate Lens
FZPA - Fresnel Zone Plate Antenna
LNA - Low Noise Amplifier
LOS - Line Of Sight Mbps - Mega bits per second
MCS - Modulation Coding Scheme
MEMS - Micro Electro-Mechanical Switches
MIMO - Multiple Input Multiple Output
PA - Power Amplifier
PCB - Printed Circuit Board
PSS - Phase Shifting Surface
QAM - Quadrature Amplitude Modulation
SIW - Substrate Integrated Waveguide
SNR - Signal to Noise Ratio
TCP - Transfer Control Protocol
TE - Transverse Electric
TEM - Transverse Electromagnetic Mode
TM - Transverse Magnetic
UDP - User Datagram Protocol
WLAN - Wireless Local Area Network

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1 INTRODUCTION

1.1 Problem Statement and Current Challenges

With the evolution of the internet and high definition video, it is no surprise that there is an ever increasing need for higher speed connections. To appease this need, the 60 GHz band has recently been made available for unlicensed communications. Countries such as Canada and the U.S. now offer 7 GHz of spectrum from 57 - 64 GHz, opening the door for massive improvements in communication speeds. Wi-Fi standard 802.11ad outlines that each of its four 2.16 GHz channels¹ will be able to obtain data rates of 6.75 Gbit/s [57] . Table 1 below compares peak communication speeds of 60 GHz Wi-Fi compared to prior protocols operating at 2.4 GHz and 5 GHz.

Table 1: Wi-Fi protocols specifications [71]

802.11 Protocol	Frequency (GHz)	Channel Bandwidth (MHz)	Channel Data Rate (Mbps)	Allowable MIMO streams
b	2.4	20	11	n/a
g	2.4	20	54	n/a
n	2.4	20	72	4
	2.4	40	150	4
ac	5	20	96	8
	5	40	200	8
	5	80	433	8
	5	160	867	8
ad	60	2160	6750	n/a

The fourth column of Table 1 outlines the data rates of each 802.11 protocol for various protocols and channel widths. These data rates are the theoretical maximums obtainable and are much lower in reality. This is mainly due to high levels of interference caused by a multitude of devices operating in close prox-

¹In some countries, there are four 2.16 GHz wide channels available in the ‘ad’ standard. Thus a single channel consists of a 2.16 GHz wide chunk of spectrum. The U.S. and Canada have only made 7 GHz of total spectrum available, therefore channel widths in these countries are slightly smaller than 2.16 GHz.

imity. The high level of interference causes an elevation in the noise floor, effectively reducing the signal to noise ratio (SNR). This forces devices to operate at lower modulation rates, which lowers throughput². The effect will be much more pronounced in standards such as ‘n’ and ‘ac’ (relative to ‘b’ and ‘g’) since they require a higher SNR to sustain their peak modulation schemes of 64 and 256 QAM respectively.

These same interference problems will be much less pronounced in ‘ad’ because of the signal’s large free space path loss and higher attenuation through walls. Thus, communications will most likely be restricted to a single room, and will not interfere with neighbouring server-client devices operating in these bands. This also implies a higher degree of security, since signals may only be intercepted by devices physically present in the same room.

The large FSPL present at 60 GHz also acts as a disadvantage by increasing radio design complexity. In order to be able to provide the power densities required to cover a room, one must have highly directional antennas, state of the art Power Amplifiers (PAs)³ and low loss circuitry.

The focus of this thesis is the design of a highly directional antenna with the ability to provide continuous coverage to a standard size room.

1.2 Target Application

The main vision for these new 60 GHz radios would be to create high speed hotspots within homes or commercial buildings. In the home, these hotspots

²The effective reduction in throughput is dependent on the level of interference, and may sometimes be enough to hamper communications completely, like in high density housing such as apartment buildings.

³State-of-the-art here refers to high gain and high linearity. The latter meaning it can accept maximal input powers without going into compression.

could be used to sync content among laptops, tablets and handsets⁴. A common example would involve syncing a 32 GB iPhone to a laptop in order to be able to more easily manage and organize pictures and videos.

Another potential application involves high definition movie downloads at airports kiosks. Downloading a 2 GB movie at home may take hours, but this can be done in 2.4 seconds using ‘ad’ as will be shown. The user would approach the kiosk and select one or multiple movies and seconds later they would be accessible on his/her tablet. Most airlines have a minimal selection of movies and small, poor resolution screens. This way, users would have access to thousands of new and old movies with the ability to watch them on a quality tablet or laptop screen.

The following sections will analyze and compare protocols ‘n’ and ‘ac’ to the new 60 GHz ‘ad’ standard in order to showcase the benefits of this new standard.

1.3 802.11n

Presently, the majority of Wi-Fi enabled devices including Access Points (APs), laptops and handsets communicate using 802.11n. Typical APs have a default channel width of 20 MHz permitting more efficient use of spectrum in areas with a high density of users. APs may also be configured to use 40 MHz channels in lower density areas, effectively doubling max throughput on a given channel. It is important to note that the data rates provided in the second column of Table 2 are theoretical maximums for single channel, and single stream, and that the actual values will always be less and depend on the level of interference.

⁴In this context, handset refers to a wireless device, such as a cellphone.

Depending on the environment, interference can be quite significant. High levels of interference result in large throughput degradation and increased wait times. In addition, the interference problem is further exacerbated because the AP is continuously transmitting during this extended period of time. Table 2 shows the resulting transfer times for throughput reductions of 5%, 15% and 30% for the 32 GB iphone sync example outlined in the previous section. A 5% reduction scenario would occur in an ideal setting where AP's are few and far apart, such as a singular household. A 30% reduction would occur in a high density areas, such as apartment buildings. These figures are approximate and are presented to give the reader an idea of the condition of current Wi-Fi protocols.

Table 2: Single channel, single stream, data rates and 32 GB file transfer times for 802.11n [71]

Channel Width (MHz)	Data Rate (Mbps)	Throughput Reduction (%)	New Data Rate (Mbps)	32 GB Transfer Time (min)
20	72	5	68.4	62
20	72	15	61	70
20	72	30	50	90
40	150	5	142.5	30
40	150	15	122.5	35
40	150	30	105	40

It is evident from Table 2 that typical transfer times are much longer than what would have been expected using the maximum theoretical data rates.

1.4 802.11ac

Protocol 802.11ac claims to provide higher data rates than 802.11n to reduce latency in big wireless transfers. However, this is not always the case; notwithstanding the added four Multiple Input Multiple Output (MIMO) channels and the ability to increase channel bandwidth, both 'n' and 'ac' are basically identical [45]. The only other parting feature is that within 10 m of the transmitter, 'n' employs 64 QAM whereas 'ac' uses 256 QAM, which allows higher through-

put. AC's use of 256 QAM is contingent upon a high enough SNR which is only possible with a direct LOS to the client. Any obstructions will reduce the SNR enough to prevent the transmitter from operating with 256 QAM and force 'ac' to revert to 64 QAM within 10 m [45].

802.11ac's advantages are truly limited at this point. The extra MIMO streams are unusable in Wi-Fi applications since most handsets are only equipped with a single Wi-Fi antenna⁵. In addition, the ability to switch to larger channels reduces efficiency in high capacity areas [45], thus in most scenarios, channel widths of 20 or 40 MHz must be used. Table 3 shows data rates and 32 GB file transfer times for varying levels of throughput reductions using the more common 20 MHz and 40 MHz channels.

Table 3: Single channel, single stream, data rate and 32 GB file transfer times for 802.11ac [71]

Channel Width (MHz)	Data Rate (Mbps)	Throughput Reduction (%)	New Data Rate (Mbps)	32 GB Transfer Time (min)
20	92	5	87	49
20	92	15	78	55
20	92	30	64.4	66
40	200	5	190	23
40	200	15	170	25
40	200	30	140	30

These numbers will be used in order to compare data rates and transfer times to the 60 GHz 802.11 ad protocol.

1.5 60 GHz 802.11ad

Table 4 compares the 'n' and 'ac' transfer times to the 6.75 Gbps interference immune 'ad' protocol.

⁵However upcoming Iphone technology will have a second Wi-Fi antenna to make use of 2x2 MIMO

Table 4: Recap of data rates and 32 GB file transfer times for all 802.11 protocols for single channel, single stream [71]

Protocol	Channel Width (MHz)	Throughput Reduction (%)	New Data Rate (Mbps)	32 GB Transfer Time (min)
n	20	5	68.4	62
	20	15	61	70
	20	30	50	90
	40	5	142.5	30
	40	15	122.5	35
	40	30	105	40
ac	20	5	87	55
	20	15	78	49
	20	30	64.4	66
	40	5	190	23
	40	15	170	25
	40	30	140	30
ad	2160	0	6750	0.64

Using a data rate of 6.75 Gbps, ‘ad’ obtains a transfer time of 0.64 minutes (38.4 seconds) for 32 GB of data. Again, this would only occur with LOS inside ten meters of the AP, but would effectively provide a speed increase of about 40x to 90x from the near range ‘ac’ protocol. Using the 60 GHz band for these types of transfers would not only reduce wait times, but also reduce interference in the 5 GHz and 2.4 GHz bands. Any system which integrated ‘ad’ compatibility would be a great complement to existing Wi-Fi technology.

1.6 Thesis Contributions

The goal of this thesis is to design an antenna for a 60 GHz radio. A successful design would allow extremely fast wireless connection speeds, made available by the high data rate outlined in 802.11 ad.

In this day and age, people are continuously transferring data over wireless networks. The benefits of this new antenna will be twofold: it will provide much faster wireless transfer speeds, as well as allow users to make full use of their in-

ternet connection speeds⁶.

This thesis outlines a switched antenna that has been designed to solve the problem of room coverage at 60 GHz. The antenna provides a high gain (18 dBi) and a steerable beam whose direction will eventually be controlled by applying a digital signal to the antenna's control ports.

This thesis offers multiple 30 GHz prototypes, which meet the gain and bandwidth requirements at 60 GHz. The design of these prototypes are easily scalable to 60 GHz.

⁶Google is coming out with 1 Gbps internet connections using fiber optics. Having routers that make full use of this bandwidth will require 802.11ad as outlined section 1.5

2 REQUIREMENTS

An elegant solution to the problem of full Wireless Local Area Network (WLAN) coverage at 60 GHz is a difficult one. Current WLAN solutions at 2.4 and 5 GHz use omni-directional antennas which have the capability to cover large spaces and propagate energy through walls with ease. Unfortunately, at 60 GHz highly directional antennas are required to overcome the massive Free Space Path Loss (FSPL), which complicates the antenna and radio technology. In addition, signal transmission through walls at 60 GHz is not possible given current indoor power density regulations [15]. This means separate APs will have to be mounted in each room requiring service. The following section details typical commercial and residential room geometries, and attempts to determine gain and beam-width requirements for the antenna.

2.1 Room Geometry

At 60 GHz, solving the problem for a particular room geometry is important due the large correlation between distance and signal loss. Referring to Table 6 in 2.2, shows that there is a 20 dB difference in FSPL between 1st and 10th meter FSPL at 60 GHz. Therefore, if necessary, the antenna gain can be relaxed for different areas of the room (zones) where the signal travels smaller distances. The purpose of this section will be to identify a high level solution on how to provide coverage to multiple different room configurations using the same unit.

The first step in such a solution is identifying a worst case scenario room. The idea is that if the unit is able to provide coverage to this room, then it will be able to provide coverage to almost any other room. This worst case scenario room was given base dimensions of ten meters by ten meters, as previous sections have used ten meters distance as a base model.

Assigning a value to the ceiling height was based on standard enterprise conference room geometries. Low ceilings have the advantage of smaller distances which result in lower gain requirements and therefore broader beams. However, with the AP so close to the floor, it must essentially provide coverage across a 180° elevation angle. High ceilings are completely opposite; longer distances but smaller angular elevation coverage requirements. Thus a “worst case scenario room” would have to solve both these problems; that is, a high gain narrow beam solution along with almost complete 180° elevation coverage.

Almost all buildings, whether residential or commercial, have ceiling heights of two to four meters. The worst case two meter ceiling scenario shown in Figure 2 is used to calculate the required angular elevation coverage⁷. Figure 3 shows a typical four meter ceiling height scenario, which provides worst case distance coverage.

Figure 2: Zone designations for ‘Room 1’ with 2 m ceilings (diagonal side view from the center of the room to the corner). This room presents the worst case scenario for angular elevation coverage.

From Figure 2, the areas of the room are cordoned off into Zones by first determining the actual room volume where requests will occur. It was deduced that requests would only be made between 0.5 and 1.5 meters from the ground since

⁷Assuming a mounting point at the center of the room. Other mounting points were considered but a central one would minimize distances.

this is the height at which the vast majority of handsets are operated, as people will either operate their handsets while seated or standing between the waist and chest level. From these baseline numbers, areas not requiring coverage are established and labeled dead Zone 1 and 2.

Figure 3: Zone designations for ‘Room 2’ with 4 m ceilings. (Diagonal side view from the center of the room to the corner). This room uses the coverage scheme outlined for Room 1 since Room 1 required the maximum amount of angular elevation coverage. Since ‘Room 2’ has the access point 2m higher than Room 1, this scenario also covers the worst case scenario path loss. Thus, solving the problem for the Zones outlined in this room would solve it for all rooms. Therefore, ‘Room2’ is the worst case scenario room.

From Figure 3, Zone 4’s beamwidth may be calculated from its gain requirement. Table 12 shows the calculated beamwidths for antennas with varying gains and aperture efficiencies. Notice that for a gain of 20 dBi, the beam-width will be 20.31° in azimuth and elevation at 100% aperture efficiency. However, such a high efficiency is unrealistic with current antenna technology. A more reasonable goal would be to attain an aperture efficiency of 75%. At this value of e_a , the beamwidth would be about 3° smaller in both planes resulting in a 17.58° beam, as shown in Equation 1. This antenna beam is very small and therefore the solution to the problem would require multiple antennas. The goal is to find a solution with a good balance between number of antenna elements and gain.

Table 5: Beamwidth vs. gain relationship at 60 GHz

Gain (dB)	e _a	BeamWidth	Gain (dB)	e _a	BeamWidth
20	1	20.31	17	1	28.69
20	0.75	17.58	17	0.75	24.84
20	0.5	14.36	17	0.5	20.28
19	1	22.79	16	1	32.19
19	0.75	19.73	16	0.75	27.88
19	0.5	16.11	16	0.5	22.76
18	1	25.6	15	1	36.11
18	0.75	22.14	15	0.75	31.28
18	0.5	18.1	15	0.5	25.54

$$\theta^2 = \frac{4\pi}{G} e_a \quad (1)$$

Where variable G is the antenna gain (linear), e_a is the antenna efficiency (%).

See example equation below:

$$\theta^2 = \frac{4\pi}{20} 0.75 \quad (2)$$

$$\theta = 17.58^\circ \quad (3)$$

Table 12 shows that at 75% efficiency with a 2 dB drop in gain, it is possible to obtain a beam that is 5° larger in azimuth and elevation. This translates to an increase in angular elevation coverage of 5.8% for two meter ceilings and 7.8% for four meter ceilings. In addition, the number of antennas needed to cover 360° in azimuth would decrease from 21 to 16. Using 16 antennas instead of 21 will facilitate design and construction due to the number's symmetrical nature. Any further reduction in the gain might result in the inability to meet the link budget, as defined in Section 2.2. Therefore, a happy medium of 18 dBi gain and 22° beam-width in azimuth and elevation is adopted as the new specification for Zones requiring this high gain.

In accordance with the new specification, Zone 4's elevation beamwidth is set to 22° . The antenna is tilted at 15° from horizontal in order to provide coverage to Zone 4 in Room 1. Figure 3 shows that the same antenna tilt would not be an optimal solution for Room 2, since the antenna in the top zone would only be responsible for a tiny section of serviceable users. On the other hand, this solution guarantees complete coverage of ten meter by ten meter rooms with ceilings ranging from two to four meters. Thus, there is only the need to manufacture one access point for two different ceiling heights, which is a huge advantage. Figure 3 now becomes a worst case scenario room, as both low ceiling and long distance situations can be covered using the same model.

As shown in Figure 3, Zone 3 has a path loss of 86 dB, which is the same path loss as Zone 4. Therefore, Zone 3 will duplicate Zone 4's solution with an added 22° tilt with 18dB gain. These two antennas will ensure reliable coverage of a 48° angular slice from horizontal.

Zone 2 provides an additional 22° but with a reduced loss of 4.6 dB from Zone 3 and 4. This allows the option of relaxing gain and increasing beamwidths, however this would require a re-sized antenna solution. In order to simplify the design as much as possible, it was decided that Zone 2 would be covered by the same antenna solution as the upper zones. The reduced path loss will result in a higher SNR at the receiver, which may be used for increased modulation rates.

As Figures 2 and 3 are displaying one half of the room, Zone 1's other 20° half is not shown, which effectively makes it one continuous 40 degree slice in elevation. As it turns out, it may be easier to develop a single antenna solution with a 40 degree beam in both cuts for coverage of Zone 1. Figure 3 shows a path loss reduction of 6.6 dB from Zone 4 to Zone 1. Using equation 1 above, the

new beamwidths can be calculated for a new gain of $18-6.6 = 11.4\text{dBi}$.

Increasing the gain by 1.5 dB to 12.9 dBi results in a beamwidth of exactly 40° which is the optimal solution. This type of coverage scenario would result in Zone 2 and Zone 1 receivers having 4.6 dB and 4.1 dB higher SNR which may permit the use of higher modulation rates.

2.2 Link Budget Calculation

This section includes 60 GHz link budget calculations for line of sight (LOS) coverage in a conference room. The main difference between a link budget calculation at 60 GHz and the common 2.4 GHz Wi-Fi band is the free space path loss. First meter FSPL is 28 dB higher at 60 GHz compared to 2.4 GHz, hence the need for high gain antennas. Table 6 below shows FSPL values for typical signal distances within a room based on equation 4.

$$FSPL(dB) = 10\log\left(\frac{4\pi d}{\lambda}\right)^2 \quad (4)$$

Where variable d is the distance between the transmitter and receiver, and λ is the wavelength. See example equation below:

$$Ex : FSPL(dB) = 10\log\left(\frac{4\pi(1m)}{(0.005m)}\right)^2 \quad (5)$$

$$FSPL(dB) = 68.00dB \quad (6)$$

As shown in section 2.1, for the case of a 10m x 10m x 4m room, the signal must travel a maximum distance of 8 meters⁸. Therefore, in order to accommodate for this worst case scenario, this link budget will use the corresponding

⁸slant length = $\sqrt{x^2 + y^2 + z^2} = \sqrt{5^2 + 5^2 + 4^2} = 8m$.

Table 6: Free space path loss at 60 GHz and 2.4 GHz

Ditance (m)	60 GHz FSPL (dB)	2.4 GHz FSPL (dB)
1	68.00	40.05
2	74.03	46.07
3	77.55	49.59
4	80.05	52.09
5	81.98	54.03
6	83.57	55.61
7	84.91	56.95
8	86.07	58.11
9	87.09	59.13
10	88.00	60.05

FSPL of 86 dB. This value, along with all other relevant radio link parameters are included in Table 7.

Table 7: Downlink link budget calculation at 60 GHz

EIRP (dBm):	
Max Output Power of CMOS Power Amp	25
Gain Transmit Antenna (Access Point)	18
Total Transmit Power (EIRP):	43
Losses (dB):	
Receiver Losses	5
Transmitter Losses	0
Polarization Mismatch	3
Atmospheric Absorption @ 8m	0.15
Free Space Path Loss @8m	86
3 dB Beam Edge Loss	3
Fade Margin	10
Integration Loss	5
Total Losses:	112.15
Signal Level at Receiver (dBm):	
EIRP	43
Gain Receive Antenna (Handset)	15
Subtract Losses	-112.15
Signal Level at Rx (Receiver Sensitivity):	-54.15
SNR at Receiver (dBm):	
Subtract Noise Floor	-81
SNR at Receiver:	26.85

The table above shows the link budget calculation for the scenario when the

AP is transmitting, called the ‘downlink’. The ‘uplink’ link budget calculation will be almost identical, with the biggest difference being a reduction in transmit power from 25 dBm to about 20 dBm. This is because in the ‘uplink’ state, wireless devices will have to make use of their battery which has limited life-time and power output. Thus, a reasonable target is thought to be 20 dBm.

The following sections will describe how each part of the link budget outlined in Table 7 is obtained, starting with Equivalent Isotropically Radiated Power (EIRP), then moving to Total Losses, and ending with Signal Level and SNR at the Reciever. Again, even though the explanations will be given for the ‘downlink’, one can assume the justifications and numbers would be identical for the ‘uplink’ with the exception of the 5 dB reductions in transmit power.

2.2.1 Equivalent Isotropically Radiated Power

The first step in a link budget calculation is to determine Equivalent Isotropically Radiated Power (EIRP). This is done by adding the transmit antenna gain to the power provided by the PA according to formula 7.

$$EIRP = P_t + G_{aT} \quad (7)$$

Where P_t is the transmit power (dBm), G_{aT} is the gain of the transmitting antenna in (dB).

Reference [11] quotes that current CMOS power amplifiers have a maximum power output of 18 dBm at 60 GHz. There has been some speculation as to whether current CMOS transistor design and manufacturing techniques are responsible for this limit or whether we are approaching the fundamental limit for output power in CMOS PA’s. The following discussion is intended to shed some light on this issue but makes no definitive conclusions on this topic.

The output power of a PA is based on voltage and current. Thus, a large output power requires these two parameters to be maximized. In PA design, voltage is limited by transistor breakdown which occurs when a critical electric field is reached. The value of this critical field is based on transistor size and more importantly, its material properties. Table 8 illustrates that silicon based CMOS transistors have a bandgap energy 10 times less than comparable technologies. This is a physical limitation of CMOS.

Table 8: Bandgap energies in different transistor technologies [6]

Parameter		Silicon	GaN	SiC
Band Gap E_g	eV	1.12	3.39	3.26
Critical Field E_{Crit}	MV/cm	0.23	3.3	2.2
Electron Mobility μ_n	$\text{cm}^2/\text{V}\cdot\text{s}$	1400	1500	950
Permittivity ϵ_r		11.8	9	9.7
Thermal Conductivity λ	W/cm·K	1.5	1.3	3.8

However, there are ways to get around this sort of physical limitation and maximize power for this technology. As mentioned previously, increasing transistor size⁹ results in a larger bandgap energy which increases breakdown voltage. There are two trends present as one makes a transistor larger:

1. Parasitic capacitance & inductance increases
2. Resistance decreases

The larger parasitics may act to increase impedance, therefore reducing current draw and PA efficiency. Larger transistors also result in a larger gate size. This means more parallel paths for the current to flow through the transistor which would decrease resistance and increase the current flow through the transistor. The design of a good PA likely involves determining an optimal sizing for the

⁹Assuming the base material remains the same. In other words, amongst silicon based transistors the only way to increase breakdown is by increasing transistor size.

transistor with regards to these two parameters.

Based on the above trends it seems favourable to make PA transistors as large as possible in order to reduce resistance and increase bandgap voltage. However, not so large as to have parasitics dominate the impedance. In practice, PA transistors are relatively large and it is not unusual for them to be thousands of times the size of other transistors inside a radio. The 18 dBm limitation may be a result of a transistor which has been optimally sized to balance between bandgap energy, resistance and parasitics. If this is the case, then power outputs hovering around 18 dBm would be a fundamental limitation of the technology.

Luckily, other techniques do exist to increase output power in PA's. One such technique involves putting multiple PA's in parallel and having their outputs add coherently. Using this method, the required output power of 25 dBm shown in the link budget calculation could likely be achieved at a higher cost.

Outside these two methods, using GaN, SiC or GaAs transistors instead of Si will vastly improve output power. In addition to having critical fields about 10 times larger than Silicon, GaN has higher electron mobility which makes it more attractive for higher frequency operation (see Table 8). It is not uncommon to see GaN, SiC and GaAs PA's with output powers of 30 dBm. However these technologies are much more underdeveloped than CMOS and thus PA's in this technology will be more costly.

It is clear that handsets at 60 GHz will be able to make use of CMOS PA's since they will only require 20 dBm output power. However, it still remains unclear whether AP's will use CMOS or another kind of technology.

The justification for a PA output power of 25 dBm and transmit antenna gain

of 18 dBi is due to the limitation on EIRP of 43 dBm [15] enforced by the FCC. The FCC also places a limit on PA output power of 27 dBm which is why such large antenna gain is required. The following sections will combine EIRP with the system losses to determine the signal level arriving at the receiver.

2.2.2 Losses

The next step is to calculate the total losses present in the system. The first two line items under the ‘Losses’ header in Table 7 is Receiver and Transmitter Loss. The main constituent of receiver loss is Noise Figure¹⁰ which can be calculated using equation 8 and 9.

$$F_{total} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_n - 1}{G_1 G_2 \dots G_{n-1}} \quad (8)$$

where:

F_1 = the noise factor of the first

element present in the receive stage

$F_2, F_3, F_n = 2^{nd}, 3^{rd}$ and n^{th} element noise factors respectively

G_1 = gain of the first element present in the receive stage

$G_2, G_3, G_n = 2^{nd}, 3^{rd}$ and n^{th} element gains respectively

$$NoiseFigure = 10\log(F_{total}) \quad (9)$$

From equation 8, noise factor of the 2^{nd} , 3^{rd} and n^{th} elements are reduced by the gains of the previous elements whereas the 1^{st} element’s noise factor is untouched. Thus, in a receive chain, the dominant factor to SNR degradation is the noise factor of the Low Noise Amplifier (LNA) since it often element ‘1’.

Reference [36] presents a TriQuint LNA with a frequency range of 57 to 65 GHz, with a 4 dB Noise Figure and a gain of 13 dB. Thus, the loss on the receive end

¹⁰Noise Figure is the overall degradation in SNR that results after the signal passes through a certain component in the transmit or receive chain. This includes insertion loss.

will be a minimum of 4 dB. Accounting for other miscellaneous losses¹¹, a final Figure of 5 dB is selected for the receiver loss in the link budget. The transmit chain is assigned a loss of 0 dB since these losses are already factored into the EIRP number.

The next loss component is antenna polarization mismatch loss. In most cellular to base station applications, a linearly polarized antenna is used at both the transmitter and receiver. Therefore, lost power due to misaligned antennas could be anywhere from zero when the antennas are aligned, to infinity when they are perpendicular. This link budget calculation assumes an average power loss of 3 dB which coincides with the antennas being tilted at 45° with respect to one another.

Atmospheric absorption loss at 60 GHz is 16 dB/km which works out to 0.016 dB/m. For a 10 meter by 10 meter by 4 meter room with the access point mounted in the center of the room, the total loss will be no more than 0.15 dB¹².

FSPL for a worst case room, as detailed in 2.1, would be at most 86 dB. This is the value that was used for FSPL in the link budget.

The next loss taken into account is the beam edge loss. This loss will vary from 0 dB to 3 dB depending on how far off the user is standing from the beam's center. A worst case scenario of 3 dB is used since the beamwidth is defined by its '3dB' points.

¹¹Transmission line losses can be considered negligible for high quality substrates and small signal distances. If the design requires significant distances to be traversed, this number will be re-visited. (See section 5.4.5 for loss calculations)

¹²slant length = $\sqrt{x^2 + y^2 + z^2} = \sqrt{5^2 + 5^2 + 4^2} = 8m$
total atm loss = $0.016dB/m * 8m = 0.13dB$

IEEE’s 802.11ad document outlines that indoor use of the 60 GHz band should include a 10 dB signal fade margin along with a 5 dB integration loss [57]. The signal fade margin arises from the signal taking multiple different routes to get to the receiver. This causes replica’s of the same signal to arrive at different times, which has the effect of spreading the pulse’s energy which effectively reduces power. Destructive interference also occurs from the time delay which again reduces signal strength at the receiver¹³. The integration loss is due to the fact that a real system is being implemented and unforeseen losses always occur. At 60 GHz the IEEE document assumes 5 dB which is what will be used in this link budget.

The total losses for this link budget amount to 112.15 dB. The next section describes how both total loss and EIRP factor into the SNR calculation for the receiver.

2.2.3 Signal Level at Receiver and Obtained SNR

For the final section of Table 7, both receiver sensitivity and SNR must be calculated. In order to calculate the signal level at the receiver, the gain of the receive antenna must be factored into the equation. According to Roland Smith, the head of the antenna research group at Ericsson Inc., “first offering commercial 60 GHz chipsets are able to achieve 15 dBi antenna gains using steerable multi-element antennas”. Adding receive antenna gain to EIRP and subtracting the total losses results in a signal level of -54.15 dBm. Comparing this to a noise floor of -81 dBm calculated below, an SNR of 26.85 dB is obtained.

$$Noise\ Floor\ (dBm) = -174dBm/Hz + 10\log(BW) \quad (10)$$

¹³Constructive interference may also occur however this is rarely the case and the net effect is always signal degradation

$$= -174\text{dBm}/\text{Hz} + 10\log(2.16\text{ GHz}) = -81\text{dBm} \quad (11)$$

The resulting 26.85 dB SNR is for a worst case distance of 8m. Nearer distances will yield higher SNRs which will allow for higher modulation rates requiring signal constellations which are more spread out. Table 9 below gives the required receiver sensitivity¹⁴ for different modulation schemes.

Table 9: Data rates and required receiver sensitivity for different Modulation Code Schemes (MCS) in 802.11 ad [57]

MCS Index	Modulation	Receiver Sensitivity (dBm)	Data Rate (Mbps)
0	DBPSK	-78	27.5
13	SQPSK	-66	693.00
14	SQPSK	-64	866.25
15	QPSK	-63	1386.00
16	QPSK	-62	1732.5
17	QPSK	-60	2079.00
18	16-QAM	-58	2772.00
19	16-QAM	-56	3465.00
20	16-QAM	-54	4158.00
21	16-QAM	-53	4504.50
22	64-QAM	-51	5197.50
23	64-QAM	-49	6237.00
24	64-QAM	-47	6756.75

The current design allows results in a receiver sensitivity of -54.15 dBm. The closest MCS index is number 20 which uses an 16-QAM modulation scheme to obtain a data rate of 4.158 Gbps. This is the type of data rate that would be attainable at the far corners of the room. As the user gets closer to the AP the receiver will monitor received power and alter its Modulation Coding Scheme (MCS). At a distance between 3 meters, the FSPL is 10.5 dB lower (see Table 6), which will result in an receive sensitivity around -44.15 dBm (MCS 24). This MCS index translates to the peak obtainable data rate of 6.75 Gbps which would be attainable between about 3 and 4.5 meter distances from the AP.

¹⁴Required SNR can easily be extracted from the table by examining the receiver sensitivity and comparing it to the noise floor.

3 TECHNOLOGY SELECTION

The room coverage solution outlined in section 2.1 requires an 18 dBi antenna with an aperture efficiency of around 75 %. Such an antenna would have azimuth and elevation beamwidths of 22° . Using this sort of antenna to provide continuous room coverage would either require a single highly steerable antenna, or the ability to switch between multiple fixed antennas, each pointed at a different section of the room. Chapter 3 investigates Fresnel Zone Plates, arrays, resonant cavities and switchable waveguides at a high level to determine their suitability for 60 GHz WLAN.

3.1 Fresnel Zone Plate Antenna

Fresnel Zone Plate Antennas (FZPA) are an attractive candidate due to their ability to provide high gain. In addition, they are inexpensive and their simple planar geometry makes them easy to manufacture using standard processes. At the heart of a FZPA is a Fresnel Lens. The Fresnel Lens was invented in 1605 by Joseph Fresnel as a method of focusing light. They were used in lighthouses as a lightweight and cheaper alternative to conventional concave refraction based lenses. The theory surrounding optical Fresnel Zones can be applied to microwave frequencies to focus energy in modern radios. This section looks at the advantages and disadvantages of using Fresnel Zones. In addition, a potential hybrid solution is analysed involving Fresnel Zones and other antenna technology.

3.1.1 Theory of Operation

The Fresnel Zone Plate Antenna can be seen in Figure 4. Lying in the XY plane is the lens which consists of alternating concentric rings made of metal and air, with a total diameter of distance ‘D’. An illuminator is also present distance ‘F’ equivalent to the lens’ focal length along the Z axis.

Figure 4: Fresnel Zone Plate Antenna [62]

The lens works based off diffraction theory. Diffraction is an interference phenomena that occurs when waves encounter obstacles. The Fresnel Lens is built by precisely locating and sizing these metal rings such that a constructive interference pattern is produced in the far field. This pattern can be predicted using the Huygen-Fresnel principle which states that “The propagation of a wave can be visualized by considering every particle of the transmitted medium on a wavefront as a point source for a secondary spherical wave” [68].

The metal rings of the lens enable you to select which portion of the incoming wavefront will be allowed to re-radiate as new-born waves. The metal rings have a dual effect, the first of which is to produce diffraction resulting in constructive and destructive interference. The second is to block the out of phase radiation to prevent any destructive interference from occurring. The following section analyses a wave as it passes from one side of the Fresnel Lens to the other, providing further insight to this design.

Figure 5 below illustrates a top view of lens at the moment when a wave is incident on it’s surface. Notice that the wavefront has a significant curvature relative to the planar lens surface, which indicates near field operation. As the curved front tries to propagate through the lens, it is interrupted by the metal

rings. These metal rings act as obstacles causing new spherical wavefronts to pass through the airy rings.

Figure 5: Top view of Fresnel Lens showing the resultant pattern of a wave passing through it. The center metal ring is labeled as ‘1’, the next metal ring as ‘2’, and so on.

Based on Huygen’s principle, the resultant wave can be calculated by treating the wavefronts passing through the airy rings as an infinite sum of point sources. For simplicity, a net pattern for the Fresnel Lens can be approximated by using a single point source per airy ring, as shown in Figure 5. The net pattern produced by side by side point sources is not a difficult one to predict. The same pattern can be observed in water waves; the pattern will contain areas with intense crests and troughs (constructive interference) which means more power is produced. However, there will be some areas where the waves completely destroy one another (destructive interference), which is undesirable for communication purposes¹⁵.

The Fresnel Lens design removes any chance of destructive interference in the far field by assuring that out of phase radiation is blocked by the metal rings at the lens surface. For a wave with a sinusoidal electric field, out of phase ra-

¹⁵It is desirable to be able to receive the signal in all areas surrounding the transceiver. Destructive interference will cause ‘dead zones’ where signals cannot be received.

diation can be defined as the section of wave that is opposing in direction, at 180° to 360°. The metal rings are placed such that the portion of the incident wave with an out of phase E-field range is reflected. Conversely, an electric field with a phase from 0° to 180° is allowed through by the airy rings.

Designing a Fresnel Lens for 60 GHz operating requires knowledge on how to size the rings properly. This requires a proper understanding of how the calculated zone sizes map themselves to a physical picture of the lens. Figures 6 and Equation 12 were used in order to calculate ring locations.

Figure 6: Labeling of i^{th} zone for use in formula 12 [22]

$$R_i = \sqrt{[F + i\lambda/2]^2 - F^2} \quad (12)$$

where:

F = focal length of the lens

R_i = Radius of the i^{th} ring (Refer to Figure 6)

λ = the wavelength

With reference to Figure 6, the first airy ring is defined as zone 1, while the first metal ring would be zone 2. Metal and airy zones would then alternate up to a

certain point, dependant on the distance of the wave source from the lens. From this Figure, it is evident that the value calculated for “R1” in Equation 12 is the radius of the first airy zone. The value for “R2” represents the distance from the origin to outer edge of the first metal ring. This results in placement of metal between “R1” and “R2”. This information is important in order to properly construct the Fresnel Zones.

3.1.2 Fresnel Lens and Hemi-Spherical Radome Hybrid Solution

This section presents a potential hybrid solution to the 60 GHz room coverage problem using a Fresnel Lens an AP placed in the ceiling’s center. The proposed solution is an array of Fresnel Lenses placed around a Hemi-Spherical Radome as shown in Figure 7. A steerable illuminator would be placed at the center of the dome with the ability to point the available energy to any of the lenses on the dome. Ideas for the feed include a planar antenna array and a Milne antenna [46], which is a dipole reflector based antenna. After further investigation, multiple issues were discovered with this approach, as outlined below.

Figure 7: Proposed FZPL coated radome [59]

The first problem with this approach is that the sphere’s cross sectional radius decreases quickly as one moves from the equator to the north pole. If one desires to make efficient use of the Radome’s surface area, then lenses must be

made smaller as one progresses toward the north pole. Smaller lenses mean smaller gains as per Equation 13. Since different room zones from section 2.1 all require the same gain, using this particular Radome lens solution is not possible.

$$G = \frac{4\pi A_{phys} e_a}{\lambda^2} \quad (13)$$

where:

G = The gain of the antenna (dB)

A_{phys} = The aperture area (m²)

e_a = The aperture efficiency (%)

λ = the wavelength (m)

Alternatively, the Radome can be resized to accommodate 21 lenses around azimuth in Lens Layer 4, the smallest circumference on the sphere. This would result in lenses in Lens layers 1, 2, and 3 being more spread out. This is an inefficient use of Radome area, therefore lenses in Layers 1, 2 and 3 can be increased in size. Either design, even at 60 GHz, would cause the physical size of the Radome to be very large. Larger domes with multiple small lenses increase the complexity of the design for the feed, and will be discussed further on in this section.

Although the Fresnel Lenses are able to provide the required gain of 18 dBi, they do so at very low aperture efficiency. Reference [63] quotes Fresnel Lens' efficiencies of only 10% to 20%, mainly due to half the power being reflected at the metal zones. Attempts have been made to increase the efficiency by using Phase Shifting Surfaces (PSS) in place of the metal zones. With this technology, the out of phase energy would be phase shifted 180° and allowed to pass through the lens instead of being reflected. The lens efficiency was able to be in-

creased to between 30% and 50% by using this method [23].

It is important to note that the efficiencies quoted above assume the lenses are properly illuminated. Proper illumination requires almost all of the power be incident on the lens face. This requires the use of a precise, highly directional and steerable feed, which is the initial 60 GHz design problem we are trying to solve. Therefore, these lenses would only complicate the design, and do not provide any added benefit.

Figure 8: Sarah Stout-Grandy’s attempts at Fresnel Zone Lens combination

The Radome based solution may be possible if lenses are made to work together. In her PhD work, Sarah Stout-Grandy attempted a couple of different lens combinations, as shown in Figure 8, in an effort to produce higher aperture efficiency. Unfortunately, the results showed no further efficiency increase. More investigation into the combination of Fresnel Lenses would be needed to rule out this approach, however it does not seem very promising for the purposes of 60 GHz WLAN.

3.2 Antenna Arrays

The next candidate for 60 GHz WLAN is planar array technology. Planar arrays are known for their ability to provide high gain with a high degree of steerability. As mentioned in previous sections, the criteria for antenna design is a gain

of 18 dBi and the ability to steer $+/- 90^\circ$ from broadside. Planar array theory will be explored to determine whether or not this technology, or a related hybrid design can meet these requirements.

3.2.1 Linear Array - Theory of Operation

To understand a planar array, it is best to first understand a one dimensional linear array. Figure 9 is presented below as a visual aid for the following explanation. This Figure shows an eight element linear array with half wavelength spacing between elements.

Figure 9: Linear antenna array [54]

To transmit in the broadside direction ($\theta = 0^\circ$), all elements must simultaneously transmit exact replicas of the pulse. A perturbation originating from each of the antenna elements will produce spherical wavelets. In the far field, the spherical wavefronts will appear flat and add constructively to produce a more intense signal in that direction.

This same principle applies to signal transmission at angles off broadside. For transmission at θ° off broadside shown in Figure 9, Element 0's pulse will be followed by Element 1 pulsing at a certain amount of time later (δ_t), and Ele-

ment 2 pulsing at $2\delta_t$. In general, for an N element linear array, each element will pulse at $(N - 1)\delta_t$ relative to the element which pulses first¹⁶. Assuming the selected angle has been mapped to a proper time delay, constructive interference will be produced in the far field in the specified direction.

Most references attempting to explain array operation refer to these time delays as phase offsets because “this is convention in the array theory world”. This is an inaccurate representation of reality and can convolute one’s understanding of how arrays work. It implies that changing the phases of signals that are all released at the same time is what produces a coherent steerable beam. This is not the case as shown in the following example.

In the scenario where end-fire transmission is required ($\theta^\circ = 90^\circ$), pulsing all antennas simultaneously would result in the nearest element’s signal appearing at the receiver first. The half wavelength separation between elements will result in each successive signal reaching the receiver a half wavelength later. Without any phase correction applied at the receiver, signals will add destructively as shown in Figure 10. However, if the transmitter applies a 180° phase correction relative to its adjacent element, the signals will all arrive as delayed versions of the initial signal as shown in Figure 11. The late arrivals of each successive pulse result in what is called “signal spreading”. This reduces the received energy per unit time, which is equivalent to reducing power. Lower received power levels translate to a higher bit error probability when compared to a simultaneous and in phase arrival.

¹⁶This assumes that the element pulsing first is labeled element 0 and that each successive element is labeled in numerical order as shown in Figure 9

Figure 10: Delay between pulses arriving at the receiver; no phase correction at the transmitter(Fractional BW ratio of 10%)

Figure 11: Fig 10 when 180° phase shifting is applied at the transmitter (Fractional BW of 10%)

The magnitude of the spread is dependent on two things:

1. The bandwidth to carrier frequency F_c ratio (aka Fractional Bandwidth)
2. The size of the array

Assuming a constant array size, if the bandwidth (BW) to F_c ratio is relatively large, the spread should not have a significant effect on communications. Figure 12 shows the scenario when the BW to F_c ratio is decreased 2.5 times. It is evident that the signal spread is not as large as in Figure 11 because the first 3.5 cycles occur during a smaller time span. An increase in array size will also contribute to an increase in spreading because of the greater distance between the first and last element. For small array sizes and small BW to F_c ratios, it is possible to implement this sort of scheme to “steer” a beam. Nonetheless, it is important to understand the effect of implementing such an antenna with phase shifts instead of true time delays.

Figure 12: For a Fractional BW of 4%

Table 10 shows that for 60 GHz Wi-Fi, the BW to F_c ratio will be about 3.33%. For the case of an 8 element linear array, the resulting pulse at the receiver would look much like Figure 12.

Table 10: Fractional Bandwidth for standard Wi-Fi bands.

Carrier Freq (GHz)	Bandwidth (GHz)	Bandwidth/Carrier Ratio (%)
60	2	3.33
2.4	0.02	0.83

3.2.2 Planar Antenna Array

The principles outlined above can be extended to a two dimensional planar array. Figure 13 shows a planar array of dipoles on the XY plane. This type of array has the ability to steer the beam along both the X and Y directions, which is an attractive property in reference to full room coverage.

There is one principal issue in using this type of antenna for indoor Wi-Fi applications; as the beam is scanned off broadside, the gain decreases with the cosine of the angle from broadside ($\cos \theta$) [13]. Placement of the array at the center of a room would cause peak gain to occur when the signal distance is the least - directly underneath the array. Conversely, the gain would be drastically reduced as the beam was steered toward end-fire where the signal must traverse the

Figure 13: Diagram of a planar antenna array [66].

largest distance. This trend makes beam steering at angles greater than $\pm 45^\circ$ [58] difficult. Using this design for worst case room coverage as detailed in Section 2.1 is not suitable.

Reference [58] outlines a potential solution for providing the extra $\pm 45^\circ$ coverage. Figure 14 shows a dielectric Radome whose material properties could be made to have a high degree of refraction near the dome's equator, and no refraction when the beam is pointed to north/south pole.

Figure 14: Dome antenna for providing an extra $\pm 45^\circ$ degree coverage for an antenna array. [58]

Although this seemed like a very attractive option, it was not pursued further than initial conceptualization phase due to the huge loss associated with 60 GHz phase shifters. Upon consultation with Carleton University's top expert on 30 GHz phase shifters Tyler Ross, a loss of 2-4dB/bit is deemed as an appropriate value for 60 GHz phase shifters. A 4 bit phase shifter would be required for 60 GHz operation in order to provide full room coverage. This would result in 8 to 16 dB of loss per phase shifter. With such massive losses, it is evident that separate transmit chains¹⁷ would be required, driving the cost of the design high. Nonetheless, some wireless companies such as Qualcomm are using arrays in their first generation 60 GHz radios. It should be noted that Qualcomm's array does not make use of this Radome design, which restricts their array to $+/- 45^\circ$ coverage.

It was decided not to pursue planar arrays because of such large loss. A more elegant, less expensive and less lossy solution is pursued in the ensuing section of this thesis.

3.3 Hybrid Cavity-Slot Antenna

In this section, a resonant cavity based approach is investigated in order to determine its suitability for indoor Wi-Fi applications at 60 GHz. A cylindrical shaped cavity is most appealing cavity option due to the symmetry of the problem at hand. A conceptual drawing of a Hybrid Cavity-Slot Antenna is shown below in Figure 15.

¹⁷A separate PA would be needed per element in the array in order to combat the phase shifter loss.

Figure 15: Left: Hybrid Cylindrical Cavity Slot. Right: Inside view of Cavity excitation mechanism [33]

The cavity consists of four main parts:

1. The metal enclosure (cavity)
2. The excitation probes shown in yellow in Figure 15
3. The slots
4. the reflectors (the black conductor shorting the top of the slot to the bottom)

This device would work by using the excitation probe to couple energy into the cavity resulting in a standing wave pattern. As long as the reflectors remained enabled, none of the energy would escape through the slots and the cavity would continue to resonate. The energy remains trapped in the cavity until one of the reflectors is disabled. The energy would then escape through the associated slot of the disabled reflector. This provides the ability to beam steer.

Since slot antennas have a small aperture, gain is usually limited to around 2-3dB [4]. Considering the large gain requirements, the need for a bigger aperture is evident. Suggestions to achieve this include mating a horn antenna directly to the slot, or mating it through a section of waveguide. Both the design of the cavity and energy extraction methods will be explained in the following subsections.

3.3.1 Cylindrical Cavity Theory

The most important aspect of cavity design is selecting the appropriate sizing for the desired wave mode. A wave mode refers to the pattern of electric and magnetic fields in a structure. Modes are either Transverse Electric (TE), Transverse Magnetic (TM) or Transverse Electromagnetic (TEM). The transverse in “Transverse Electric” means the electric field is aligned perpendicular to the direction of wave propagation. The same follows for TM modes where the magnetic field is perpendicular, and TEM modes where both electric and magnetic fields are perpendicular to wave propagation. These designation are suitable for most propagating waves, including waves in free space and waveguides.

Wave modes in cavities SEEM to be mislabeled in reference to the basic definition of TE, TM and TEM. A good example of a POTENTIALLY mislabeled propagation mode was found in reference [47], which uses Figure 16 to illustrate the ‘TM010’ mode.

Figure 16: TM 010 mode as illustrated in [47]

This Figure shows an electric field in the z direction and a magnetic field in the ϕ direction. The Poynting vector tells us that $E \times H$ should produce the direction of propagation of the wave. Crossing the z vector with the ϕ vector gives a wave propagation in the radial direction. In this scenario, it is evident that both the

electric field and magnetic field are perpendicular to wave propagation. Based on elemental wave mode definitions, this particular mode should be classified as *TEM* and not *TM*.

However, there are multiple problems with the above train of thought. The first is that the modes inside a cavity are called ‘resonant modes’ and not ‘propagation modes’. There is never a ‘propagating wave’ inside a cavity, and therefore, the Poynting Theorem does not apply. Secondly, waveguides cannot support TEM modes and thus, the above rational is incorrect in both instances. The ‘TM010’ mode is in fact the correct labeling of the field configuration shown in the cylindrical cavity of Figure 16.

Regardless, it is not of great importance what the modes are called, so long as they can be reproduced. Before reproducing resonance in the simulator¹⁸, it is important to determine which mode should be activated. Equation 14 below dictates cavity sizing for a given resonant frequency and mode.

$$f_{nml} = \frac{c}{2\pi\sqrt{\epsilon_r\mu_r}} \sqrt{\left[\frac{p_{nm}}{a}\right]^2 + \left[\frac{l\pi}{d}\right]^2} \quad (14)$$

where:

a is the cavity radius (m),

d is the height (m),

p_{nm} is the nm^{th} order bessel function, depending on the mode (see Table 11)

$\epsilon_r\mu_r = 1$ for free space,

$l = 0$ or 1 , depending on the mode (0 for TM_{010})

c is the speed of light in a vacuum

¹⁸Empire is the simulator being used in this thesis. It is introduced in Chapter 4

Table 11: p_{nm} values for TM_{0n0} modes [47]

TABLE 12.1 Parameters of TM_{0n0} Modes

Mode	k_n	ω_n
TM_{010}	$2.405/R_0$	$2.405/\sqrt{\epsilon\mu} R_0$
TM_{020}	$5.520/R_0$	$5.520/\sqrt{\epsilon\mu} R_0$
TM_{030}	$8.654/R_0$	$8.654/\sqrt{\epsilon\mu} R_0$
TM_{040}	$11.792/R_0$	$11.792/\sqrt{\epsilon\mu} R_0$

Re-arranging for ‘a’ we obtain:

$$a = \frac{c}{2\pi\sqrt{\epsilon_r\mu_r}f_{nml}}p_{nm} \quad (15)$$

For a frequency of 60 GHz and selecting mode TM_{010} with $p_{nm} = 2.405$ from Table 11, a cavity radius of 1.9 mm is obtained. This cavity sizing is too small for two reasons: firstly, the final prototype would have many slots cut out around its circular face for multi directional use (about 16) and the current circumference would restrict the number of slots. Secondly, since the gain needs to be boosted, a larger aperture is required as mentioned previously. The closer the slots are, the less room there will be to flare out to a bigger aperture.

In order to compensate for size requirements, Equation 15 was re-examined to determine how the cavity could be made larger. Since variable p_{nm} increases with second subscript ‘n’, Equation 15 is re-computed for TM_{040} . The new cavity radius is 9.3mm which should prove large enough to fulfill both aperture size and spacing needs.

A 9.3 mm cavity was constructed and simulated in an attempt to excite mode TM_{040} . The mode pattern is shown in Figure 17. As illustrated by the legend in the top right corner, red represents areas of intense electric field and blue represents least intense electric field.

Figure 17: Simulated mode patten of TM_{040} in 8.05 mm cylindrical cavity

This cavity was built using a quarter wavelength probe located at the center of the cavity much like Figure 15. Through simulation, optimal radius was resolved to be 8.05 mm; slightly less than the 9.3 mm predicted by theory.

3.3.2 Extracting Energy from Cavity

In an attempt to extract energy from the cavity, a single resonant slot was cut out from the cavity face. The slot was made resonant at 60 GHz and placed with its length oriented perpendicular to the electric field direction. A multitude of different slot sizes and locations were simulated with no success.

Reference [4] gives an example of a simulated slot antenna with fractional bandwidth of 8% at 2.45 GHz which would result in 4.8 GHz of bandwidth at 60 GHz. Again, simulation results produced nowhere close to this. One potential explanation for this may be that only a small fraction of the energy interacts

with the slot causing poor performance.

Another option investigated involved using a loop structure shown in Figure 18 to extract energy. The amount of energy that couples onto the wire is proportional to the amount of magnetic field lines intersecting with the loop's face. Optimum performance could only be obtained if the loop were extended toward the center of the cavity. This structure is difficult to manufacture using standard processes, therefore ruling out the option of using cylindrical cavities for 60 GHz WLAN.

Figure 18: Loop coupled cavity [33]

3.4 Guided Wave Structure

As mentioned earlier, the antenna solution at 60 GHz has to be low loss, highly directional, and easily manufacturable. It also must be able to provide complete room coverage for a worst case scenario room. Guided wave structures are attractive because they meet all of these requirements. In the following chapter, an in-depth investigation of a switchable waveguide structure theory is analysed for 60 GHz operation.

4 A SWITCHABLE WAVEGUIDE

A 3D rendering of a switchable waveguide is shown in Figure 19¹⁹. It works by using a central probe to excite a metal enclosure defined by the orange vias, along with two ground planes completely covering the top and bottom surfaces (not shown in the Figure). Once the energy is present inside the structure, it will propagate until it encounters the four pairs of reflectors shown in blue. Each reflector pair is digitally controllable to be in an enabled or disabled state. If three reflector pairs are enabled, each pair will reflect the incoming energy and its only escape will be through the pair of disabled reflectors. In essence, the result is a programmable antenna whereby emission direction can be selected via digital signal. Manufacturing such a structure requires the highly evolved fabrication techniques associated with PCB technology.

Figure 19: End to end design of 4-way switchable waveguide. Not visible in the Figure are the 4 horn antennas and the top and bottom metalization for ease of viewing.

The design procedure begins with the sizing of the metal enclosure and associated waveguide sections. Manufacturing a basic metal enclosure in the PCB

¹⁹All the various features shown in Figure 19 will be dissected in detail in the coming sections.

world requires a special approach; the most common of which is a technology called SIW short for Substrate Integrated Waveguide. Section 4.1 introduces this technology and escorts the reader through basic sizing of the SIW.

The second design step of the SIW is the investigation of the heart of the design: the controllable reflectors, also known as the switches. Section 4.2 describes the factors effecting switch performance. In addition, different reflector configurations are tested in order to determine which arrangement produces peak performance.

In Section 4.5, the various performance metrics of diodes and Micro Electro-Mechanical Switches (MEMS) are compared to determine which technology is most suitable as a control block for the reflectors.

In Section 4.6, the design of the dielectric to air transition from the SIW to the horn antenna is described. To complete the design, section 4.6.2 contains the theoretical design and simulation of the physical horn antenna.

In order to conform with the M.A.Sc. time constraints, it was decided to build prototypes at 30 GHz. The design at 30 GHz is completely scalable to 60 GHz; all performance metrics obtained at 30 GHz are representative of those at 60 GHz. The prototype designs should comply with the antenna specifications of 18 dBi with 22 ° beams as outlined in Chapter 2. As 7 GHz²⁰ of bandwidth is being used at 60 GHz, creating a required fractional bandwidth of 12%. Using this percentage, at 30 GHz this translates to a requirement for a bandwidth of 3.5 GHz.

²⁰57-64GHz, see Section 1

4.1 Substrate Integrated Waveguide Theory

At this point, there might be some confusion pertaining to exactly what is meant by a ‘metal enclosure’. The particular design shown in Figure 19 forms its metal enclosure by the intersection of four 30 GHz waveguides. Multiple other designs exist, but are always made by intersecting multiple waveguide sections as shown.

The first design step is to select waveguide dimensions for the target frequency. The standard waveguide class ‘WR-28’ has an operating range between 26.5 GHz and 40GHz with a standard width dimension of 7.1 mm and height of 3.55 mm. It is important to note that this is the sizing specifically for an air filled waveguide. Since SIW requires a substrate, its relative dielectric constant will have an effect on the waveguide sizing. The substrate selected is a Teflon based Rogers material 5880 with dielectric constant 2.2 and loss tangent 0.0009. The low dielectric constant was selected to maximize wavelength according to Equation 16.

$$\lambda = \frac{c}{\sqrt{\epsilon_r}f} \quad (16)$$

where:

λ is the wavelength (m)

ϵ_r relative permittivity of the material

f is the frequency of operation (Hz)

c is the speed of light in a vacuum (3.0×10^8 m/s)

At such high frequencies, the minuscule sizes begin to challenge current PCB manufacturing capabilities, and manufacturing tolerances begin to have a significant effect. The larger the wavelength, the easier it is to have a reliable design manufactured. The new waveguide dimensions for a teflon filled waveguide are calculated using the following three equations.

For a rectangular waveguide, the cutoff frequency f_c is,

$$f_c = \frac{c}{2\pi\sqrt{\epsilon_r\mu_r}} \sqrt{\left(\frac{n\pi}{a}\right)^2 + \left(\frac{m\pi}{b}\right)^2} \quad (17)$$

where:

c is the speed of light in a vacuum (3.0×10^8 m/s)

ϵ_r relative permittivity of the material

μ_r relative permittivity of the material

m and n are dependant on the propogation mode

a is the width of the waveguide (m)

b is the height of waveguide (m)

Most waveguide applications choose to propagate the dominant TE_{10} mode to prevent degenerate modes from oscillating. Energy contained in different modes will travel at different velocities. This results in signal dispersion or pulse spreading. This pulse spreading can result in intersymbol interference which increases the bit error rate, effectively degrading communications. This is normally a concern in media such as optical fibre where the signal must traverse long distances, however distances traversed in this antenna will be short enough for this to be a non-issue. The second problem with propagating degenerate modes is loss. If multiple modes are propagated, separate probes in different places are required to capture the energy on the receive side. Since the design aims for low complexity, only one probe is present to capture energy. This will result in un-captured energy which is seen as a loss. Thus the design must only propagate the dominant mode TE_{10} .

For $n=1$ and $m=0$ the expression simplifies to,

$$f_c = \frac{c}{2\pi\sqrt{\epsilon_r\mu_r}} \left(\frac{\pi}{a}\right) \quad (18)$$

Re-arranging for waveguide width ‘a’,

$$a = \frac{c}{2\pi\sqrt{\epsilon_r\mu_r}} \left(\frac{\pi}{f_c} \right) \quad (19)$$

All parameters are known except the cutoff frequency, f_c . Reference [44] recommends operating a waveguide from 125% to 189% of the cutoff frequency. The lower limit is placed to prevent operation too close to f_c while the upper limit is placed to prevent oscillation of degenerate modes.

To accommodate this restriction f_c was chosen to be 20 GHz. This produces an acceptable band of operation from 25 GHz to 37.8 GHz. This is plenty of spectrum for the target of 3.5 GHz bandwidth centered at 30 GHz. Substituting $f_c = 20\text{GHz}$, $\epsilon_r = 2.2$ and $\mu_r = 1$ into Equation 18 results in a width ‘a’ of 5 mm. This new width is 70% of the width calculated for an air filled waveguide; the impact of the dielectric on size is significant. As seen in Equation 18, the height of the waveguide ‘b’ has no bearing on guide operation for the TE_{10} mode. Therefore, a width of 5 mm will be used as a benchmark to size the SIW.

In SIW technology, the waveguide walls are formed using many closely spaced vias. The top and bottom walls are metal layers on the PCB. It is important to know how large and how far apart vias should be spaced before beginning the design. In order to minimize energy leakage, it is usually best to reduce via spacing as much as possible. Unfortunately, minimal distances are usually governed by manufacturer tolerances. In this case, manufacturing restrictions require that vias be spaced a minimum of 10 mil from the edge of a via to the edge of another. With this in mind, the formulae below are used to select SIW via parameters [43].

$$p < 2d \qquad d < \frac{\lambda_g}{5} \quad (20)$$

Figure 20: Via diameter (d) and spacing (p) for SIW technology [43]

Where ‘d’ is the via diameter, ‘p’ is the via spacing from center to center, and the guide wavelength, λ_g , is given below,

$$\lambda_g = \frac{2\pi}{\sqrt{\frac{\epsilon_r(2\pi f)^2}{c^2} - \left(\frac{\pi}{a}\right)^2}} \quad (21)$$

With $\epsilon_r = 2.2$, $f = 30$ GHz and $a = 5$ mm, equation 21 gives: $\lambda_g = 9.1$ mm or 358 mil. Substituting this result into Equation 20 gives a maximum via diameter of d ; 1.82 mm or 71.6 thou.

In order to approximate the straight edge of a waveguide wall, it seems favourable to minimize the diameter of the vias. A via diameter of 20 mil restricts via spacing to less than 40 mil. For a minimum edge to edge spacing of 10 mil, via separation will be 30 mil, which gives a 10 mil margin from the minimum spacing requirement. To summarize, final sizing will be a spacing ‘p’ from center to center of 30 mil and a via diameter of 20 mil. The next section examines the design of the switchable waveguide’s main feature: the switch.

4.2 Switch Theory

The element that gives the waveguide structure its switching capability is a monopole reflector. This section uses Empire simulation software to see how different switch configurations behave and then attempts to intuitively explain the behaviour observed. Before continuing, it is important to further introduce Empire.

Empire is a software tool which facilitates the process of solving Maxwell's four equations governing all electromagnetic phenomena. It allows users to build up complex problems in three dimensions using simple shapes with varying electric and magnetic properties. The tool is one of the fastest electromagnetic solvers currently on the market which is what makes it so attractive. All simulation data presented in this thesis has been obtained through Empire. Figure 21 illustrates a side view of a waveguide containing a monopole reflector built in Empire.

Figure 21: Single disabled monopole reflector

The dashed lines labeled P1E and PC2 are ports in the simulation software used to measure S-Parameters. In red are the waveguide walls, and in orange are two blind vias. The left via is the waveguide excitation probe. The right via is the controllable reflector which is sized to be $\lambda_g/4$ high.

The grey lines represent an air cut out in the metallic ground plane. In the case of the excitation probe, this cut out allows for energy to flow in and out of the waveguide. In the case of the reflector, the cut outs allow for a switch such as transistor, diode or a MEMS to be placed between the reflector and the ground plane. This gives the ability to control whether the reflector is floating or grounded based on the signal applied to the switch. This is important

because it is what gives the waveguide the ability to be directionally reconfigurable.

When the switch is off, there is no connection between the ground plane and the via. The result is a non-resonant quarter wavelength conductor which allows energy to pass by. The result from simulation can be seen in Figure 22 and shows $> 10dB$ match over a 12GHz bandwidth from 25 to 37 GHz.

Figure 22: Disabled single reflector s-parameters

However, when the switch is turned on, there is a connection made between the ground plane and the via. As a consequence of image theory²¹, the conductor now appears as a resonant half wavelength reflector and perfectly reflects incoming energy. The results from simulation for the SIW environment show an optimal post height of about 50 mil which is just under $\lambda_g/6$. However this post height does not account for the added length needed to short the post to the ground plane. This extra distance adds about another 10 thou to the reflector length bringing it to 60 thou or $\lambda_g/6$. The discrepancy between $\lambda_g/4$ and $\lambda_g/6$ post sizing will be explained further on in this section.

²¹The effect of image theory is to double the apparent length of the conductor when it is connected to a ground plane.

Figure 23: Enabled single reflector s-parameters

As it turns out, the grounded reflector does a very good job at reflecting energy as shown in Figure 23. Between 29 and 31 GHz only 10% (-10dB) or less of the energy is not reflected. This results in a fractional bandwidth of 6.67% when the reflector is in the enabled state.

In an attempt to increase the enabled state bandwidth to match the 11.67% required at 60 GHz, different switch configurations were postulated and tested.

Figure 24 below illustrates another switching style which has two in-line monopole reflectors.

Figure 24: Disabled double monopole reflector

Initially, it was thought that the optimal spacing between the two reflectors would be $\lambda_g/4$. The rationale was that energy reflected from the second post would add in constructively with energy reflected from the first post. If the posts truly operated in this fashion, then a separation of a half wavelength should produce destructive interference, and the result would be little to no signal being detected at the input. It turns out that this method of thinking does not accurately predict the outcome. In reality, very good performance is still observed with half wavelength separation. Therefore, a new understanding of how this sort of post configuration behaves is sought out.

In his book Foundations for Microwave Engineering, Robert E. Collin has this to say about capacitive posts²²

“The post (...) behaves more like an LC series network connected across a transmission line. When the depth of penetration is between $0.7b$ and $0.9b$, it becomes resonant and acts almost like an ideal short circuit. For lengths greater than this resonant length, the post is equivalent to a shunt inductive susceptance...” Collin continues, “there is no way to assign a unique inductance or capacitance to this obstacle since there is no unique way to define the voltage, current, or characteristic impedance for a waveguide. Nevertheless, this obstacle

²²Known in this document as ‘reflectors’ or just ‘posts’

does produce a reflected wave with a reflection coefficient that is correctly determined in terms of the equivalent normalized reactance and susceptance of the obstacle as connected into the equivalent-transmission-line model of the waveguide.”

As this excerpt points out, an alternative way to think about this complex behaviour is to think of each post individually as a series ‘LC’ circuit separated by a length of transmission line. The resulting equivalent circuit is shown in Figure 25.

Figure 25: Equivalent circuit for two posts separated by some amount of transmission line

Parameters C_{dx} and L_{dx} represent the equivalent inductance and capacitance present from the added length of t-line²³. This circuit can be further simplified as shown in Figure 26.

Figure 26: Simplified equivalent circuit for two posts separated by some amount of transmission line

The net result is a circuit which comprises of two different resonant LC circuits.

²³Equivalent T-line circuit is built assuming a low loss environment such that $G=0=R$

The location of each resonance can be controlled by adjusting a multitude of parameters such as post separation and the physical size of the posts. As Collin points out, there is no way to assign values to post inductance and capacitance. The above only serves as a means to intuitively think about the behaviour of these posts. Accurate post sizing and spacing results were obtained using simulation data. This LC model approximation is reinforced with the results obtained in the following simulation results.

4.3 Switch Configuration Selection

As mentioned in the previous section, optimal results were obtained for a post height of $\lambda_g/6$. An optimal spacing of $\lambda_g/5$ was determined after a sweep of post separation. The results for a double reflector arrangement with these parameters are simulated and shown below in Figure 27 below.

Figure 27: Enabled dual monopole reflector simulation results

The graph shows a large improvement in the “enabled state” bandwidth rela-

tive to the single reflector. The bandwidth has increased from 2 GHz to 5 GHz which results in a fractional bandwidth increase from 6.67% to 16.67%. One interesting feature present in Figure 27 is the double resonance with nulls occurring at 29 GHz and 31 GHz. This feature will be analyzed and discussed further in Section 4.4.1.

The double reflector arrangement was also simulated in the “disabled state” as illustrated in Figure 28. The results are almost identical to the single disabled reflector shown in Figure 24, providing about 12 GHz of bandwidth from 25 GHz to 37 GHz. Thus, the double reflector proves to be a much better option due to increased “enabled state” bandwidth and an unchanged “disabled state” bandwidth from the single reflector option.

Figure 28: Disabled dual monopole reflector simulation results

A triple reflector configuration was also simulated and produced a triple resonance as expected, however the overall performance was no better than the double reflector. Therefore, the following sections will be dedicated to investigating

parameters associated with the double reflector. Nonetheless, most of the theory and results are transferable to the other reflector configurations since they operate on the same principle.

4.4 Double Reflector Switch Design

As discussed previously, the double reflector option was chosen for this design. The following subsections demonstrate simulation results, including post spacing, post height, waveguide height, cut out size, post radius, effect of via pads, and post location. The simulation setup in Figure 24 of section 4.2 was used to carry out these simulations. Since it is a closed system, the inspection of either S_{21} or S_{11} will suffice to judge the performance of the switch. In other words, the power that is not observed at one port must be present at the other.

4.4.1 Post Spacing

This section examines the effect of post spacing on enabled and disabled state reflector performance. Varying this post spacing allows one to control the frequency separation of double resonance in the enabled state. When the reflectors are physically closer, the resonances get further apart. Conversely, as the reflectors are moved further apart, the resonances get closer together. This trend is presented in Figure 29 for post spacings around $\lambda_g/5$. The effect of this variance on the disabled state performance was negligible which is why the trend has not been shown graphically.

Figure 29: S21 vs. Reflector separation for an enabled pair of reflectors at different post spacings

Figure 29, along with those to come, specify the post parameters in thousandths of an inch or ‘mils’. Table 12 is included as a conversion guide between feature size and percent wavelength. The values are calculated by dividing the feature size (in mil) by the guide wavelength of 358 mil calculated in Section 4.1.

Table 12: Feature sizes translated to percentage of one wavelength

Feature Size (mil)	Fraction of Wavelength (λ)	Feature Size (mil)	Fraction of Wavelength (λ)
30	0.0837	100	0.279
35	0.0977	110	0.307
40	0.111	120	0.335
45	0.125	130	0.363
50	0.139	140	0.391
55	0.153	150	0.418
60	0.167	160	0.446
65	0.181	170	0.474
70	0.195	180	0.502

4.4.2 Post Height

Another important parameter of reflector design is the height of the reflectors.

In this section, three different trends are investigated in the following order:

1. Variation of both post heights equally
2. Variation of post 1 height while keeping post 2 height constant
3. Variation of post 2 height while keeping post 1 height constant

Only enabled state operation for these variations are presented here since reflector height has little to no impact on disabled state operation. It is important to note that these variations were done with all other parameters held constant. Figure 30²⁴ illustrates the effect of varying both post heights together around values just below $\lambda_g/6$.

Figure 30: Effect of changing both reflector heights equally in the enabled state, 40mil to 50mil

As intuition would predict, increasing the height enables the reflection of lower frequency signals. Conversely, decreasing the height allows reflection of higher frequency signals. Another noticeable relationship is the bandwidth increase as

²⁴It is important to note the small portion of the curve where the S21 goes above 0 dB. Since the antenna does not have a gain element, this is impossible. The reason for this is because the simulation time was cut short in an effort to reduce wait times. This caused a spur in the response which shot the S21 over 0dB. The effect of shortened simulation times was determined individually for each different model by comparing the shortened version to the regular version. The differences between the responses were negligible in almost all cases which allowed for the use of these shortened simulation times. There are a couple plots in coming sections which have small portions of the S21 residing over 0dB.

post heights get larger. This trend can be seen further in Figure 33 below as post height is increased beyond 50 mil.

Figure 31: Effect of changing both reflector heights in the enabled state, 50 mil to 60 mil

A bandwidth of 3.5 GHz at a center frequency of 31.5 GHz is observed for a post height of 40 mil. In comparison, the 60 mil post provides a bandwidth of 7 GHz at a center frequency of 26.5 GHz. Thus, a large bandwidth advantage presents itself if operation at a lower frequency is an option. The exact physical mechanism responsible for this bandwidth increase is speculative at this point. Nonetheless, knowing this trend exists and how it behaves suffices to move forward with the design.

Next, the effect of altering the height of reflector 1 while keeping reflector 2 constant was examined. Reflector 1 is reduced below the equal height point of 50 mil, in 5 mil increments. The result of this experiment is shown in Figure 32. Decreasing the post's height decreases its inductance which modifies the resonant frequency of that post. The effect is that the leftmost resonant dip remains in place while the rightmost dip is moved higher in frequency.

Figure 32: Effect of decreasing reflector 1 height in the enabled state

In addition, a decrease in height produces a larger bandwidth at the expense of poorer reflective capability. Figure 32 shows that 99% of the power being reflected over a 6 GHz bandwidth for the 50 mil post. In comparison, the 40 mil post reflects 90% of the power over a bandwidth of 7 GHz. This results in 9% decrease in reflected power but an increase in fractional bandwidth of 3.3%. It is up to the user to decide which configuration to use depending on whether there are larger requirements on bandwidth or power.

The height of reflector 1 was also increased above the equal height point of 50 mil to 70 mil in 5 mil increments. The result from simulation is shown below in Figure 33.

Figure 33: Effect of increasing reflector 1 height in the enabled state

The effect produced is similar to what occurred in Figure 32. As the height is increased, the leftmost resonance dip is dragged down to lower frequencies which acts to increase bandwidth. A slight degradation is observed in reflective capability, however it is nowhere near as pronounced as when reflector height is decreased. For 99% power transfer (-20dB) the 50 mil post produces a bandwidth of 4 GHz in comparison to 5.5 GHz bandwidth for the 60 mil post. The actual bandwidth improvement by the 65 and 70 mil posts are not obtainable from this simulation setup because the waveguide prevents operation below 23 GHz. Nonetheless, it appears as though the bandwidth continues to increase with only slight degradation of reflective capability. This trend provides antenna designers working in this technology with yet another degree of freedom to meet requirements.

The same simulations are performed for a height sweep of reflector 2. Figure 34 is a plot of S21 as the height of reflector 2 is decreased below 50 mil to 30 mil in 5 mil increments. The response is almost identical to that produced for a height

reduction of reflector 1, shown in Figure 32. Again, shortening the reflector results in the rightmost resonance dip being pulled up to higher frequencies which makes the structure broader band. Unfortunately, the additional bandwidth produced is accompanied by a large deterioration in reflective capability represented by the higher S21.

Figure 34: Effect of decreasing reflector 2 height in the enabled state

Figure 35 illustrates the trend as the secondary reflector height is increased above the equal height point. Again, the result is very similar to the scenario where the height of reflector 1 is increased, as shown in Figure 33. Significant bandwidth advantages are available, however they are accompanied by the deterioration of reflective capability.

Figure 35: Effect of increasing reflector 2 height in the enabled state

The deterioration of S21 is worse when manipulating reflector 2 as compared to reflector 1. Therefore, it is more beneficial to extend the first reflector if this sort of tuning is required.

In summary, there are advantages and disadvantages to leveraging any of the three aforementioned design modifications. Firstly, for same height reflectors, it was shown that altering their heights shifted the center frequency. For this specific design, shorter reflectors were detrimental to bandwidth, so it is better to lengthen them if possible.

Secondly, reflectors with different heights were explored. It was found that any significant reductions in either reflector height relative to the other degraded performance. However, lengthening either reflector resulted in greater bandwidth without much effect on reflective ability. The front reflector did produce less S21 degradation. If a larger bandwidth is required, lengthening the front reflector is the best option.

The main disadvantage to having different reflector heights is that it requires an additional board layer. This adds cost, fabrication complexity and increases the likelihood of manufacturing error. None of these are typically major issues, therefore if the added bandwidth is required, this is a very good option. For manufacturing simplicity, posts of the same height were used in this SIW design.

4.4.3 Waveguide Height

This section explores the effect of waveguide height on the structure's reflective and transmissive capabilities. Figure 36 shows simulation results for a waveguide height of 2 times, 2.4 times, and 2.8 times reflector height.

Figure 36: Reflective capability for varying waveguide height (2x, 2.4x, 2.8x) - enabled state

As height is increased past double the height, the rightmost resonance dip is pulled down to lower frequency, resulting in decreased bandwidth. The response begins to degrade significantly as height is increased further, as shown in Figure 37.

Figure 37: Reflective capability for varying waveguide height (3x, 3.2x, 3.4x) - enabled state

Any increase above 3x results in reflectors becoming progressively less effective in the enabled state. At 3.4x (or 170 mil), the reflectors are only good for 90% power reflection over a 2 GHz bandwidth. This is a huge drop in performance over the 2x height where 90% power reflection is obtained over 8 GHz.

Optimal enabled state performance occurs at a height of 100 mil or double the reflector height. However, this height does not provide optimal disabled state operation. Figure 38 shows the switch's performance in the disabled state for the same waveguide height increments.

Figure 38: Effect of waveguide height on disabled state operation

It is apparent from the Figure that maximum power transfer occurs between 2.6x and 2.8x. Selecting a height of 2.7x produces optimal disabled state operation while also providing a 6 GHz bandwidth in the enabled state. Such a large bandwidth provides a good margin over the requirement of 3.5 GHz. A height of 2.7x translates to 135 mil which will be the benchmark waveguide height for the design.

4.4.4 Size of Cutout

Another design parameter under investigation was the size of the cutout around the posts. In the enabled state, the short between the pad and the ground plane is constructed differently in the prototype design²⁵ unit than it would be in the final ideal product. The differing structure of the shorts effect how the cutout behaves electromagnetically and therefore different configurations should be considered.

The prototype design has a grounding mechanism like that shown in Figure 39.

²⁵This thesis constructs the prototype version.

In this case, a short circuit is formed during manufacturing by leaving a small section of the ground plane in tact between the pad and the ground plane. This sets the reflector to the enabled state, and is not reconfigurable.

Figure 39: Top view (left) and side view (right) of grounded reflectors in the prototype phase

This is different than the structure of the final ideal design which would mate a PIN diode between the via pad and the ground plane, as shown in Figure 40. This would allow the design to be reconfigurable, as these diodes could be controlled digitally, allowing reflectors to be enabled and disabled at will.

Figure 40: Side view of grounded reflector in an ideal final design using a PIN diode

It is important to note that since pads are so tiny (18 mil to 21 mil diameter), a switch with a small package will be required. Maacom [38] has a commercially available diode in a package size of about 4 mil wide by 9 mil long by 2 mil tall,

with small pads²⁶. It is rated up to 40 GHz, thus this diode seems like a very attractive candidate.

The grounding mechanism used in Figure 39 was used for the following simulations in this section. For all of the following tests, the cutout sizing was varied by changing the radius of the cut from the center of the via to the edge of the cutout.

In order to quantify the effect of cutout sizing on switch performance, the size of the cutout was decreased from 14 mil to 8 mil in 2 mil increments for a constant post radius of 5 mil. The result is shown in Figure 41.

Figure 41: Effect of changing reflector radius in the enabled state for a cutout radius of 17.5 mil

Reducing the size of the cutout shrinks the gap between the conductor and the bottom wall of the waveguide. This effectively shortens the post, and increases its resonant frequency. This increase in center frequency is accompanied by a large degradation in reflective capability. The exact cause of this degradation

²⁶More exploration on pad sizing will be covered in Section 4.4.5

is hard to pinpoint. It seems as though it is more difficult for any RF energy that couples onto the post to make its way onto the ground plane as the gap is smaller.

This phenomenon could be occurring simply because the ground plane is approaching a changing magnetic field. The closer the ground plane is to the post, the more intense the field will be, possibly inducing excess noise on the ground plane. By assuming that the simulator accounts for situations such as the one described above, the design process was continued.

The size of the cutout was then increased from 14 mil to 24 mil. As expected, the larger gap effectively increases the post length which causes it to resonate at a lower frequency. As evidenced in Section 42 reflectors with longer lengths promote bandwidth increases, which is apparent here. In addition a degradation in S21 is seen which is likely due to poorer match.

Figure 42: Effect of changing reflector radius in the enabled state for a cutout radius of 17.5 mil

With regards to disabled state operation, the cutout has no effect on performance with regards to the sizes simulated. After running multiple simulations

around cutout sizing, it is apparent that it is not an exact science. As a general rule, it is safer to size gaps on the larger side when in doubt. In addition, for larger posts, the cutout radius should be increased accordingly in order to maintain a constant gap size. For this prototype design, cutout sizes of 17 mil for smaller radius posts and 24 mil for larger radius posts were used.

4.4.5 Post Radius

This section details the investigation of the effect of post radius on the switch's reflective and transmissive capabilities. Figure 43 illustrates the results as post radius is varied for enabled state operation with all other parameters held constant. In particular, the cutout radius is held constant at 17.5mil. Therefore, by increasing post radius, we are decreasing the air gap between the post and edge of the cutout.

Figure 43: Effect of changing reflector radius (4 mil to 1 mil) in the enabled state for a constant cutout radius of 17.5 mil

This plot reveals 3 trends that are present when post radius is increased: ²⁷

²⁷Most current pcb shops have minimum hole sizes (diameter) of 5 mil so anything below a

1. Bandwidth increase
2. Degradation in S21
3. Response is shifted to lower frequency

It is perhaps helpful to identify how fatter reflectors change the response. As per discussion with the head of the Antenna Research Group at Ericsson Inc., Roland Smith, the first trend on the list is related to an increase in post surface area. As this is where most of the current travels, the larger surface area provides more ‘lanes’ for current to flow up and down the post’s surface. Since the path that current takes in a conductor isn’t always linear, the gap from the smallest lane length to largest lane length will probably be wider. This allows the conductor to host lower frequency currents, which results in the BW increase in the lower frequency direction, as per Figure 43.

A confident explanation for exactly what causes line items 2 and 3 is still speculative. However one potential reason for the second trend would be a shift in the impedance of the post. Regardless, simply recognizing the presence of the trends and how to alter them is enough to continue with a design. Thus, based on these trends, a designer can choose a higher bandwidth environment if a lower center frequency can be accepted along with a slight decrease in power reflection.

Figure 44 demonstrates the trend as reflector radius is further increased in the enabled state.

radius of 2.5 mil would not be manufacturable

Figure 44: Effect of changing reflector radius (4 mil to 16 mil) with a constant cutout radius of 17.5 mil in the enabled state

Again, increasing post radius increases bandwidth, degrades reflective capability and shifts the center frequency lower. However, for a certain cutout size, there is a limit to how much the radius can be increased, as discussed in Section 4.4.4. As previously stated, once the edge of the conductor gets too close to the ground plane, the switch fails as evidenced by the 16 mil simulation in the figure above.

These simulations were performed with a constant size cutout of 17.5 mil, which only leaves a 1.5 mil gap in the 16 mil post case. Increasing the gap size between the via pad and the ground plane will permit the use of this larger radius post. All in all, the large bandwidth advantage presented by increasing the post radius outweighs the reduction in power reflection, since it is very hard to get anything better than 90% (10dB) power reflection in practice. However, even if a large post radius seemingly improves things in the enabled state, this may not be the case for the disabled state. The results for a sweep of post radius in the disabled state are shown in Figure 45.

Figure 45: Effect of changing reflector radius on disabled state performance

The graph shows that larger reflectors worsen performance in the disabled state. However their impact on performance is much less significant than what was observed in the enabled state. Thus, in a design, it is better to gravitate toward larger posts for the enabled state bandwidth advantage.

The last trend to be investigated here involves varying the radius of only one of the reflectors. Figure 46 shows the result as the size of the front reflector is varied.

Figure 46: Effect of changing front reflector radius in the enabled state

The response is fairly one dimensional in the sense that a fatter front reflector increases enabled state bandwidth with almost no side effects. Before any further examination is done in the enabled state, the same sweep is performed for a post in the disabled state shown in Figure 47 below.

Figure 47: Effect of changing front reflector radius in the enabled state

The trend is similar to that which occurs when both reflector heights are swept at the same time; that is, a fatter post reduces power transmission. For this design, a post width of 13 mil for a 17.5 mil cutout is the critical value at which impedance appears to be infinite after the first post, preventing any power from passing through.

As it turns out, the response is identical when an identical sweep is performed by varying the radius of the back reflector. In terms of manufacturing, the varying a single post diameter is an ideal tuning parameter, as no additional fabrication complexity is needed. This is a huge advantage as compared to varying a single post height, as mentioned in Section 4.4.2.

4.4.6 Effect of Pads

Standard PCB manufacturing techniques involve the addition of pads to the top and bottom of any via as shown in Figure 48.

Figure 48: Top view (left) and side view (right) of grounded reflectors in the prototype phase

The pads are actually advantageous in switchable waveguide design for two reasons. The first is that they act to increase bandwidth, and secondly, they are able to shift the response's center frequency. This parameter is easily varied during manufacturing and thus, provides the designer with an extra parameter to tune the design.

Figure 49 illustrates the trend in the enabled state as both top pads on the reflector posts are increased in size. The pads sizes are quoted in terms of via overhang. For example, if a via is 5 mil in diameter, to produce a 2 mil overhang, the pad should be 7 mil.

Figure 49: Effect of changing size of top pads on both reflectors in the enabled state

Larger top pad sizes effectively lengthen the post, which results in the S21 curve being shifted to lower frequencies. The larger pad sizes also act to increase bandwidth. The physical mechanism responsible for this bandwidth increase is very similar to that responsible for bandwidth increase in fatter posts described in Section 4.4.5. During disabled state operation, no effect was observed for pad size variations in the same range. Moreover, conducting the same set of simulations on the bottom pads had no noticeable effect on reflective or transmissive capabilities. Thus, antennas requiring more bandwidth may be obtained by enlarging the top pad radius without any side effects.

From a manufacturing standpoint, a major advantage associated with varying a pad size is simple and can be done with multiple PCB variations within the same panel²⁸. Altering this design parameter to tweak a response is much more convenient than using for example, post height variations, since it does not re-

²⁸A pcb panel consists of multiple substrate and metal layers ‘glued’ together. The size used for production of the switchable waveguide was a 12” by 18” panel in which about 40 variations were done.

quire different substrate heights.

4.4.7 Location of Posts

In the basic simulation model used throughout this entire section, keeping the spacing constant and altering the location of the posts relative to the feed has no effect on performance. However, in versions of the designs where multiple of these switching junctions are integrated, their locations matter. Their location impact antenna matching and performance. These impacts will be discussed further in Section 5.

4.5 Diodes and Micro Electro-Mechanical Switches

This thesis only deals with the switchable waveguide structure and does not attempt to integrate a configurable control line structure. Nonetheless, diodes and MEMS are investigated in order to determine their suitability for providing switchability to the posts in an ideal final design.

After a good deal of research, it became apparent that low loss, high switching speed diodes are scarce at 30 GHz, not to mention 60 GHz. The PIN diode in [38] seems to be the best available presently, and is quoted for use up to 40 GHz. However, reference [29] makes use of the diode in its 60 GHz reflect-array which is very promising! The diode's equivalent RLC circuit in the 'short circuit' state is given below in figure 50.

Figure 50: Diode equivalent RLC circuit

The impedance for this circuit is derived in [28] and the result is given below:

$$Z_{in} = \frac{R}{1 + \omega^2 C_p^2 R^2} + j \left(\omega L_s - \frac{\omega C_p R^2}{1 + \omega^2 C_p^2 R^2} \right) \quad (22)$$

Using values of 0.5 nH for L_s , 28 fF for C_p and 4 Ω for R, the formula can be simplified to:

$$Z_{in} = Z_{eq} = R + j\omega L_s \quad (23)$$

A plot of real and imaginary parts of Equation 23 is given:

Figure 51: Diode impedance vs. frequency

As evidenced by the equation and Figure 51, this diode's real resistance is a constant 4Ω and the imaginary impedance varies linearly with frequency. At the frequency of interest of 30 GHz the impedance is $4 + j94.2\Omega$. This results in an impedance completely dominated by susceptance, giving an impedance magnitude given of about 94Ω as shown below.

$$|Z| = \sqrt{|4|^2 + |94.2|^2} \quad (24)$$

$$|Z| = 94.28 \quad (25)$$

Such a large impedance will kill current flow through the diode; and any current that is induced, will suffer a significant loss. In order to use the diode at this frequency one must minimize this impedance. The susceptive component can be removed entirely by placing a capacitor with reactance $-j94.2$ in series with the diode as shown below:

Figure 52: Diode model with series capacitor (C_s) added in order to decrease diode impedance

The value of C_s at 30 GHz is calculated as follows.

$$Z_c = \frac{1}{j\omega C} \quad (26)$$

$$-j94.2 = \frac{1}{j\omega C_s} \quad (27)$$

$$C_s = 56.3 \text{ fF} \quad (28)$$

The resulting impedance of the circuit in Figure 52 is 4Ω - purely real. This new diode model is simulated on its own in the conducting state for various values of C_s around 50 fF and compared to that of an ideal short circuit.

Figure 53: S21 for varying values of series capacitance C_s

Based on these results, for optimal performance centered around 30 GHz, a C_s value of 42 fF looks optimal. This would result in a maximum loss of 0.5 dB per reflector inside a 7 GHz bandwidth²⁹. At first thought, this loss may seem large since multiple reflectors would be used in a final solution. However, this is not necessarily the case as the following explains.

Figure 54 below shows how one may estimate the loss for a waveguide with two separate outputs (2-way setup) with one reflector per branch. A similar results would be produced with two reflector per branch, thus the above scenario will suffice to obtain a ballpark figure for loss in the manufactured 2-way structure outlined in Section 5.4.3.

²⁹7 GHz at 30 GHz translates to a fractional bandwidth of 23.3% which is well above the target.

Figure 54: Switch loss associated with the diode for a 2-way. The cross hatched circle indicates the disabled reflector.

In this setup, the right reflector is enabled, thus it should apply a 10% power loss. A total output of 1 W by the central feed results in 0.5 W incident on each reflector. This results in a loss of 0.05 W at the right reflector and 0.45 W being reflected toward the left branch. Meanwhile, the other half of the power (0.5 W) from the feed immediately propagates down the left branch. Since the left reflector is disabled, this energy will experience no loss through the diode. This will produce a total of 0.95 W of power at the output of the left branch. Therefore, only 5% (0.22 dB) of the power is consumed which is half of the 10% that was expected per reflector. In essence, this is due to the parallel combination of diode resistors halving the total resistance.

A 4-way reflector setup is also manufactured as shown in Figure 69 of Section 5.4.4. A top view is shown here in Figure 55.

Figure 55: Switch loss associated with the diode for a 4-way. The cross hatched circle indicates the disabled reflector.

Here $1/4^{th}$ of the power is incident on each of the posts. If a total of 1 W is output by the feed, then 0.25 W is incident on each reflector. Since each reflector incurs a 10% loss, the actual power reflected will be 0.025 W less per enabled reflector. For the three active posts, this loss totals 0.075 W which is a loss of 0.34 dB.

The switches above only have one post per branch to simplify calculations, but the outcome for two posts per branch would be similar. Total loss for a 16-way directional antenna (designed with multiple 2-way or 4-way configurations) would be around 0.9 dB ($0.3 \text{ dB} * 3 \text{ junctions}$) assuming the energy encounters 3 switching junctions, as shown in Figure 56.

Figure 56: Potential 16-way directional switchable waveguide for continuous azimuth room coverage

Another consideration to take into account is the switching speed of the diode - for this PIN diode, switching speed is 5ns (0.2GHz)- and whether it can accommodate switching speeds required in 802.11ad Wi-Fi.

The required switching speed in the 802.11 ad protocol is based on the size of an empty User Datagram Protocol (UDP) or Transfer Control Protocol (TCP). When a server-client communication session is entered in a Wi-Fi setting, any requests made by the client have to come in the form of packet structured according to the above protocols. Minimally, a communication must consist of at least one packet. So, in theory, it is possible for multiple client devices to initiate one packet conversation with the AP one after the other. The amount of time it takes for the AP to get the information across to the first client will determine whether diode switch speed or packet transfer is the limiting factor to

AP client delegation speed. The calculations below are sourced from [60] and shows packet empty packet sizing for TCP and UDP.

TCP

Ethernet frame - 24 Bytes

IPv4 Header (without any options) - 20 bytes

TCP Header (without any options) - 20 Bytes

Total Size of Empty TCP Datagram - $24 + 20 + 20 = 64$ bytes

UDP

UDP header - 8 bytes

Total Size of Empty UDP Datagram - $24 + 20 + 8 = 52$ bytes

A worst case packet size of 52 bytes is equivalent to 416 bits. The transmission of this packet at 802.11ad peak data rate of 7Gbit/s results in a packet transmission time of 59 ns. This is one order of magnitude away from the minimum switching time of 5 ns. For MCS indexes with lower modulation schemes, the time cushion from the switch's minimum switching time would only increase. Therefore the 802.11ad data rate is the limiting factor in this scenario.

Another technology which could be considered is the use of MEMS. These have the advantage of having extremely low resistance on the order of 0.5Ω and a fast enough switch speed in the nano second range. The main issue with MEMS is their lifespan hovers around 1 billion cycles. The average handset or access point may switch around 1 billion times per month which renders current MEMS technology unusable for this application.

4.6 Dielectric to Air Transition

Another challenge involved in the design of such an antenna is the transition from the dielectric filled waveguide (or PCB) to air. This transition was done in two steps. The first step was concerned with the transition of dielectric filled waveguide to an air filled waveguide. The second step involved transforming the impedance of the air filled waveguide to that of a horn antenna in free space. Coincidentally, horn antennas turn out to be a great candidate for 60 GHz WLAN since they have high gain, high aperture efficiency, and low loss.

The following section will examine the design of the transition from dielectric to air filled waveguide.

4.6.1 Dielectric Filled Waveguide to Air Filled Waveguide

The main two aspects of the transition design are dielectric taper and waveguide flare angle. No simulation results are presented in this section due to the number of design parameters of this structure. Instead the author attempts to provide the reader with an intuitive understanding of how each element functions and how they work together.

The dielectric taper is required because abrupt changes in dielectric constant result in large signal reflections. A gradual dielectric taper acts to slowly change impedance from that of one dielectric to another much like the foam absorbers in an anechoic chamber.

Various dielectric taper configurations were simulated but the most effective turned out to be tapering in only the H field plane³⁰. Figure 57 shows a top view of the taper. The point was left 40 mil thick since this is as small as could

³⁰The H plane is the plane perpendicular to the posts

be manufactured without physical deterioration.

Figure 57: H plane dielectric taper (PCB top view), where the cross hatched pattern indicates a disabled post.

There is not much flexibility in selecting dielectric taper angle since the waveguide width has been designed to be 173 mil and the minimum feature size for the taper is 40 mil. The only parameter that can be varied is the dielectric taper length or distance from the start of the taper to the end of the taper. For larger dielectric taper lengths, the transition is more gradual, therefore the response improves for longer lengths.

Next, different waveguide flare angles were simulated for constant dielectric taper lengths. The waveguide flare was also only in the H plane; that is, the width was progressively made larger. The results showed a range of flare angles with very low amount reflections (-20dB).

To summarize, for dielectric taper, a longer tongue provides a more gradual impedance transformation and thus is generally a better choice. In terms of waveguide flare, for this 30 GHz design one should target a final width of somewhere around 280 mil which is the width associated with WR-28 (26.5 GHz to 40 GHz). There will likely be an optimal flare angle range for the particular dk taper selected, thus the designer will have some flexibility. The next

section examines the design of the horn antenna responsible for transforming the impedance of the air filled waveguide to that of free space.

4.6.2 Horn

Horn antennas are an effective means of propagating energy from a waveguide structure out into free space. As mentioned above, they act as an impedance matching tool to transform the waveguide impedance to that of free space. They also satisfy the required criteria of being low loss, high gain, and have high aperture efficiency. With aperture efficiencies from 40% to 80% [70], the horn can provide the required gain while maintaining a small physical size.

There are a few important design parameters to keep in mind before proceeding. In the final design, each horn will be responsible for covering a certain section in azimuth. It is desirable to minimize the number of horn antennas needed to cover a room in order to reduce complexity and loss. More elements results in more switching junctions and a larger feeding structure (PCB), both of which account for almost all of the loss associated with the switchable waveguide design.

Reducing the number of horn antennas involves designing for larger beamwidths. However, since high gain (18 dBi) is required, the only way to achieve larger beamwidths without reducing the gain is to increase the aperture efficiency³¹. This can only be done by lengthening the horn for a constant aperture size [70]. Using an initial horn length of 2 inches with an aperture of 1" by 1", aperture efficiency was examined for slightly different aperture sizes, flare angles. The finalized design was a horn antenna measuring 1.73 inches long with an aperture measuring 950 mil in the E plane by 1271.5 mil in the H plane. For a beamwidth

³¹Equation 29 illustrates this relationship

of 22° in azimuth and elevation with a gain of 18 dBi, aperture efficiency is found by rearranging the equation below.

$$G = \frac{4\pi}{\theta_{el}\theta_{az}}e_a \quad (29)$$

Where θ_{el} is the angle in elevation (radians), θ_{az} is the angle in azimuth (radians), and e_a is the aperture efficiency (%). For this scenario, $\theta_{el} = \theta_{az}$, and gain is 18 dBi (63 linear), which reduces the equation to:

$$e_a = \frac{G\theta^2}{4\pi} = \frac{63(0.384)^2}{4\pi} = 0.74 \quad (30)$$

An aperture efficiency of 74% is obtained, which is near the practical maximum of 80% efficiency for horn antennas. This is ideal for a 60 GHz Wi-Fi solution which requires high gain and large beams. The simulated radiation patterns for the elevation and azimuth cuts are shown below in Figure 58 and Figure 59 respectively.

Figure 58: Antenna elevation pattern ($\phi = 0$)

Figure 59: Antenna azimuth pattern for ($\theta = 90$)

These results show gain hovering around 18 dBi and 22° beamwidths over a 5 GHz (17% fractional BW) bandwidth. These results are directly inline with the requirements outlined in Section 3. Therefore, a horn antenna is a good option for the 60 GHz switchable waveguide and will be used for the prototyping of this design.

5 DESIGN METHOD & RESULTS

Chapter 4 gave the reader insight on how the individual components of a switchable waveguide functioned. However, no guide was given on how to put all of these pieces together. The first part of this section gives the reader a general guide on how to assemble the individual parts presented in Chapter 4 to create a switchable waveguide. The second part of this section presents simulation results and testing results for each of the three switchable waveguide prototypes manufactured. Many different versions of each of the three different prototypes were simulated, manufactured, and tested; however for brevity, only models producing the best results are included here.

5.1 Design Process

Chapter 4 outlines design methodologies for individual aspects of the design. The aim of this section is to complete the design path by meshing individual design aspects together to obtain a complete model for prototyping. The following provides the reader with chronological outline of the design steps and decisions to be made when carrying through with the design of a switchable waveguide:

1. Choose Specs - center frequency, BW and gain
2. Select substrate and PCB metalization
3. Determine SIW parameters (width, height, via diameter, spacing)
4. Select connector (obtain or create simulation model)
5. Size excitation probe & reflectors
6. Build simulation model
7. Attach connector in step 4 to waveguide excitation probe
8. Simulate structure and ensure good S11 and S21 performance in enabled and disabled states

9. Add dielectric to air waveguide transition (ensure good S11 and S21)
10. Add horn antenna (ensure good S11 and S21)
11. Intersect two 1-ways to make a 2-way, or four 1-ways to make a 4-way.
12. Create a 16-way by combining blocks in step 11

The first step is to select specs for the switchable waveguide to adhere to, such as frequency, bandwidth, gain and aperture efficiency. With switchable waveguide technology, the user has control over all of these parameters. Frequency is a function of waveguide and via sizing. Bandwidth is mostly limited by enabled state reflector bandwidth. The author's intuition figures this bandwidths between 10% and 25% might be attainable for optimized 8-way or 16-way antennas. The gain and aperture efficiency are simply a function of horn antenna dimensions. In practice, gain and aperture efficiency are limited by device sizing requirements.

Next, the designer should select the PCB substrate; at high frequencies, the permittivity should be made as low as possible to maximize wavelength, and by extension, maximize PCB feature sizes. This will make the design more reliable³² and make the manufacturing process easier. In addition, because of the increased loss at high frequencies, a substrate low loss tangent should be selected.

PCB metalization should also be selected. This design used 1/4 ounce copper which results in a 2 mil metal layer thickness. Since skin depth is on the order of microns at these frequencies, the metal thickness is large enough to prevent energy leakage. In addition, thinner metal layers provide increased etching and drilling accuracy which helps minimize error.

³²Tolerances won't be as much of an issue.

Once the designer knows the substrate permittivity, guide wavelength can be calculated and used to size all the relevant SIW parameters. These parameters include waveguide width, height, and wall via diameter and spacing.

Following waveguide sizing, the designer should select a connector appropriate for the frequency range selected. The prototypes designed in this thesis used a coax to PCB surface mount connector³³. This step is important in order to obtain a connector model to be able to continue running accurate simulations. Although the connector is quoted as 50Ω , this is not necessarily the case; especially across such large bandwidths. Thankfully, Petra Microwave owner Peter Frank had a model for this connector on hand, hence the decision to select it. Attempts to obtain other connector models from other microwave connector companies were unsuccessful due to I.P. issues.

Steps 5, 6, 7, 8 can all be done concurrently as the following explains. In traditional waveguide designs, the feed is sized to half the waveguide height ‘b’ or around $3/16\lambda$. However this number is not concrete and can be adjusted to optimize the specific design. To determine its exact height, the designer should fabricate a simulation model consisting of five things: the SIW enclosure, the ground plane cutouts, the vias, the reflector(s) and the connector model as shown in Figure 60. An ideal coax excitation port should be attached to the connector model, and an ideal waveguide energy sink³⁴ port should be used to terminate the waveguide. Then, parameters such as feed height, location and diameter should be swept to determine an optimal sizing.

³³This was a 2.4 mm DC-50 GHz Molex connector. Part number: 733870020

³⁴A sink port here refers to a waveguide port which is perfectly matched and absorbs all incident energy.

Figure 60: Potential simulation model

Once optimal enabled and disabled state performance is obtained for the model above, one can move to steps 9 and 10. Step 9 requires the addition of the dielectric filled waveguide to air filled waveguide transition to the simulation model. The base design for dielectric taper and waveguide flare can be tweaked in order to maintain optimal enabled and disabled performance.

Finally, step 9 involves the addition of the horn antenna. Including horn antennas in simulation models are extremely disadvantageous for two reasons. The first is that simulating models with horn antennas results in much longer simulation times since it is multiple times larger than the PCB. Secondly, it is much more difficult to obtain S21 curves in the simulator if the energy is allowed to radiate out from the horn face. For these reasons, it is desirable to be able to design and optimize a full structure without the horns and then just add them in as a last step. This works out for the design of the 1-way, but unfortunately, the design of more complex switchable waveguides appear to require the horn antennas to be attached for accurate results. This was not totally confirmed but appeared to be the case. It is definitely worth spending time at this stage of the design process attempting to develop a model which produces the same s-parameter results with and without the horn antenna.

The above outlines the design from start to finish of a 1-way switch. To design 2-way and 4-way switches, the designer should intersect two or more 1-way switches and merge the feeds to obtain a base model. Once the new model is created, the designer should begin varying parameters such as post height, spacing and diameter. A new parameter introduced in this case is post location. Post location now matters because one reflector pair is effectively acting as the shorted end of a waveguide. It turns out that the distance of the feed from the backwall is a relatively sensitive parameter and therefore reflector distance from the feed is crucial. This adds an extra dimension to the design and optimization of multi-way switches. Apart from this, multi-way switch design is exactly the same as the 1-way and involves establishing a base model and varying parameters in the simulator and observing their effect of performance.

The methodologies described above were executed to design 1-way, 2-way and 4-way switch blocks. Other attempts included 3-way and 5-way junctions, however proper standing wave patterns could not be created.

The 16-way switch required for a 60 GHz WLAN was built by connecting multiple 2-way blocks to a central 4-way switch like that shown in Figure 56 of section 4.5. The figure shows two differently designed 2-way junctions: an L-shaped junctions and a T-shaped junctions. This thesis only dealt with the design and manufacture of the T-shaped junction. The associated L-shaped junction and overall 16-way structure were simulated at 60 GHz and produced a 90 % power transmission over a 7 GHz bandwidth to each of its 16 horn antennas. It is important to note that this simulation was just a first order approximation and did not include any of the following: pads, cutouts, Dk transitions, SIW vias, connector model. The model also did not include any attempts to optimize the individual switching junctions. Therefore, in the author's estimation, this is very promising! In the following section, testing results for each of the three proto-

types are presented starting with the single switch configuration AKA “1-way”.

5.2 Simulation Results

The whole aim of this section was to attempt to maximize enabled state bandwidth since this was always the limiting factor to a better design. The main mechanism for improving this bandwidth is to use different reflector heights or diameters. However, both of these options were removed in order to simplify things. The former was removed in order to keep the board to two layers. The latter was removed in order to reduce the number of design variables. In hindsight, this was not a smart move as the varying of this parameter has a direct impact on increasing enabled state bandwidth, and by extension, overall device performance. Nonetheless, great results were still obtained even with the removal of these two key design parameters. Future switchable waveguide designs will easily be able to top performance metrics quoted here.

5.2.1 Single Switch Configuration

The purpose of this section is to present simulation results for the single switch configuration. Figure 61 below illustrates a 3D rendering of the final simulation model of this version. At the top right of the screen, in light grey is the horn antenna, followed by the tapered dielectric to air transition leading to the PCB. The PCB contains the SIW walls in orange, along with the excitation probe (bottom) and reflector posts in blue. In black, we see the conductor pads, and the grey represents the ground plane cutouts.

Figure 61: Final PCB design of single switch configuration with 1 layer of vias

After a lot of trial and error with the simulator, the results in figure 62 were obtained. The graph shows an enabled state S_{11} of about 0 dB from 26 GHz to 33 GHz. This represents a fractional bandwidth of 23% which is much higher than expected. The results are just as impressive for a disabled state bandwidth of over 9 GHz. However, there has been concern over what portion of the disabled state power is leaked through gaps in the design versus emitted through the horn antenna. The simulator is not able to produce S_{21} curves for this scenario but it does have a tool which allows monitoring of the power incident on any of the 6 planes enclosing the model. The model can be setup such that any leaked power from the gaps is prevented from attaining the same plane being irradiated by the horn antenna. This is done by adding a metal sheet which extends from the outside of the waveguide at the throat to the boundaries setup in the simulator. This effectively shields the plane monitoring the horn's output from all sources of energy. This allows for the realization of the S_{21} curve shown

in figure 62.

Figure 62: Single Switch Configuration (1-way) S-parameters

An S21 hovering very closely to 0 dB is observed between 27 GHz and 35 GHz. It is important to note that using this method of obtaining S21 does not account for mismatch loss. The monitoring planes only tell you what percentage of the radiated power is incident on each the simulation model's boundaries. If little to no power is radiated this will not be an accurate representation of the S21. However, if over 90% of the power is radiated, this method is considered to be a good approximation. This is why monitoring the S21 depends on the value of S11!

The other trend present in Figure 62 is the wide band enabled state performance reflected by an S11 around 0 dB from 26 GHz to 33 GHz.

Figure 63 and 64 show the simulation results for the azimuth beam and the elevation beams.

Figure 63: Simulation results for 1-way azimuth cut

Figure 64: Simulation results for 1-way elevation cut

Since the antenna was simulated without losses, the y-axis of the graph represents antenna directivity (dBi). This particular 1-way antenna hovers around 18 dBi to 19 dBi with beamwidths around 22° in both cuts as required. Another noticeable feature is the sidelobes present from 10 dB to 12 dB down in the elevation cut. These are of no real concern to any part of the design, however serve as a sanity check when comparing testing results later.

5.2.2 Double Switch Configuration

The purpose of this section is to present simulation results for the dual switch configuration. Figure 65 below illustrates a 3D rendering of the final simulation model of this version³⁵.

Figure 65: Final PCB design of dual switch configuration with 1 layer of vias

After a lot of trial and error with the simulator, precise sizes for all features were determined for optimal performance. These sizes produced the results shown in figure 66. The graph shows most of the power escaping through port 2 which is desirable. Optimal operation appears to occur within a 4 GHz range between 28 GHz and 32 GHz where the energy lost to radiation via port 3 appears to be between 1.3 dB and 0.4 dB. This translates to a fractional bandwidth of 13.3% which is above the requirement of 11.7% required by 60 GHz WLAN.

³⁵This figure has been altered from its original state and is not to scale. The length from the center of the split to the edge of the board has been shrunk down in order to provide a complete view of all features.

Figure 66: 2-way S-parameters

It is evident that a significant amount of power (8% to 18%) leaks out the other port. Potential reductions here could be achieved by adding a third post with an equivalent LC made to resonate at 28 GHz. Other methods also include choosing different reflector widths or heights as outlined earlier. This could improve bandwidth and reduce the amount of power leaked. These options should be explored in later versions of the design.

The simulation results for the 2-way elevation and azimuth radiation patterns are shown below in Figure 67 and Figure 68 respectively.

Figure 67: Simulation results for 2-way azimuth cut

Figure 68: Simulation results for 2-way elevation cut

A clear reduction in gain (almost 3 dB) is seen at 28 GHz due to the loss through port 3. This is not ideal and this unit should be re-designed to improve this. However, testing results tell a much different story; in a good way, as will be seen.

5.2.3 Quadruple Switch Configuration

The quadruple switch configuration was the final prototype to be built and tested.

Figure 69 illustrates a 3D rendering of the final simulation model.

Figure 69: Final PCB design of quadruple switch configuration with 1 layer of vias

After a lot of trial and error with the simulator, precise sizes for all features were determined for optimal performance. These sizes produced the results shown in Figure 70.

The graph displays an S21 curve hovering around 0 dB which indicates that almost all the power is escaping through the disabled port 2. Even though the data is not shown here, the S21 remains around 0 dB for about an extra 0.25 GHz on either side. This results in a fractional bandwidth of 11.7% exactly which matches the requirement of 11.7%. Again, improving this value could easily be achieved by choosing different heights or diameters for reflectors within a pair or even adding a third post. There are many options!

Figure 70: 4-way S-parameters graph

Figure 71 and Figure 72 illustrate the azimuth and elevation beamwidths respectively. Again, most of the gain reduction occurs at 28 GHz which coincides with the edge of the acceptable band of operation. However the results here are much improved over those observed in the 2-way. All directivity values hover between 17.5 and 18.5 dBi with 3 dB beamwidths in both cuts of around 22° as required.

Figure 71: Simulation results for 4-way azimuth cut

Figure 72: Simulation results for 4-way elevation cut

The coming section discusses manufacturing considerations which is then followed by the presentation of the antenna's test results.

5.3 Manufacturing

After obtaining good performance simulation results, the PCB's were manufactured by Accurate Circuit Engineering (ACE) in California. There are a few considerations with respect to PCB manufacturing techniques that have not yet been mentioned which will be covered here.

Since the prototype boards were all two layer, a bond film was required. The bond film is a small layer of glue placed in between PCB layers to hold them together. Ideally, a bond film with Dk 2.2 and minimal thickness would be used to glue the two layers together. However, the only stock ACE had on hand was a Rogers 2929 bond film with Dk 2.9 which comes in 1.5 mil thicknesses. Since the pads were etched on 2 mil thick copper, two bond film layers ($1.5 + 1.5 = 3\text{mil}$) were required.

In order to maximize etching and drilling accuracy, the ground layer thickness was minimized. Quarter ounce copper (2 mil thick) was used since it was the thinnest the PCB shop could do.

The PCB were mounted on an aluminum base and aligned with the horn antennas using alignment pins. This ensured the horn antennas and PCB's were almost perfectly aligned. Using only screws would have provided up to 6 mil of wiggle room in either axis and this was seen as too much during the design process. However, the PCB's were tested with just screws and the performance was almost identical. Future generations might consider removing alignment pins all together however, but they can't hurt.

Figure 73: Pins on left and right aligning PCB to horn antennas

The standart coax to PCB compression fit connectors worked fine for the single and double prototypes (modified version of the connector shown in Figure 74). Screws came up through the bottom of the PCB, grabbed the connector and secured it tightly to the board to make a good connection with the excitation via. However, this did not work for the 4-way prototype because of two reasons. The first is that the screws responsible for mounting the connector interfered with

the waveguide structure. Secondly, the connector footprint actually intersected with pairs of disabled vias which would have shorted them to ground and turned them into enabled reflectors. Therefore, the connector had to be modified in two ways. Firstly, a custom piece was designed as an unintrusive alternative to the screw mounting technique. This piece is shown in Figure 76. Secondly, slots were cutout in the connector to prevent shorting of the disabled vias shown in Figure 74. The imprint left after the connector was mounted is shown in Figure 75.

Figure 74: Modified coax to PCB connector with four slots cutout

Figure 75: Imprint left on connector after securing it to the board

Figure 76: Part designed to compression fit connector base to PCB

5.4 Testing Results

5.4.1 Measurement Setup

The test setting was Ericsson's anechoic chamber. The chamber's frequency range was initially unknown, however, from the dimensions of the absorbers, one may extract the frequency at which the anechoic chamber was designed. The absorbers had about a 3" by 3" base and a height of around 5". Reference [67] cites that the absorbers function best when the absorber height is $\lambda/4$. A height of 5" translates to height of about 12.54 cm. Therefore, the absorbers function best at $4 \times 12.54 = 50.16$ cm. This is equivalent to a frequency of 600 MHz. Since the performance of an anechoic chamber is governed by the lowest test frequency, absorption at 30 GHz will still be high. Therefore, this chamber is suitable for these testing applications. The equipment outlined below was used to test the antennas:

- 2 x DC to 50 GHz cables (3dB loss / cable) - W.L. Gore & Associates - Part Number: 0Z0CJ0CK0360
- 1 x 10 MHz to 50 GHz Network Analyzer - Keysight PNAx N5245A
- 1 x coax to WR28 connector - unknown
- 1 x 2.4 mm female to 2.92 mm female microwave connector - unknown
- 1 x 24 dBi WR28 gain standard - A-info
- 1 x Electronic Calibration kit (ECal) - Agilent N4693-60001

Figure 77 shows a test setup with the probes in both the gain standard and Device Under Test (DUT) aligned horizontally to perform an elevation cut on the 1-way antenna. The analyzer is shown with one cable emanating from port 1 which is attached to the transmitting gain standard (left). The cable on the right is connected between the receiving DUT and port 2 of the PNAx. Both transmit and receive antennas were each stacked on styrofoam amounting to

a height of ten inches. The height was chosen such that the antenna's beams would not intersect with the wooden base and underlying table; thus preventing reflections from interfering with results.

Figure 77: Elevation plane cut

Using this test setup, four plots were produced for each antenna, namely, S11, radiation patterns (azimuth and elevation), cross polarization isolation, and gain vs. frequency.

The equivalent of a device S11 plot was obtained using the test setup illustrated above by recording S22. To ensure test reliability, this measurement was performed with the DUT's active aperture rotated to point directly at one of the chamber's walls.

The radiation patterns in azimuth were measured using the test setup shown in Figure 78. This test setup is different from Figure 77 in that the probes are now both aligned vertically. To obtain the pattern, the aperture centers of the antenna's are first aligned. Then, the DUT is rotated from -45° to $+45^\circ$ in 2.8125° increments. For each position, the S21 is recorded at six frequencies sur-

rounding 30 GHz, each spaced 1 GHz apart.

Figure 78: Azimuth plane cut

For the elevation cut, the DUT was turned on its side as in Figure 77, and the aperture centers of the tx and rx antennas are re-aligned. Once proper alignment is obtained, the DUT is rotated again, and the S21 is recorded for 6 frequency points around 30 GHz for each 2.8125° angular step from -45° to $+45^\circ$.

The cross polarization plot is obtained by comparing the S21 of the antenna in the co-pol position to the S21 in the cross pol ³⁶. Only a single measurement of cross polarization isolation was done for a tx antenna probe oriented at 90° relative to the rx antenna's probe and the antenna faces directly aligned. This was seen as sufficient since cross pol isolation will only get better with rotation from direct alignment.

³⁶Taking a cross pol measurement involves probes to be oriented at 90° from one another

Finally, a gain vs. frequency plot was obtained by measuring antenna S21. Obtaining the gain of the antenna is slightly more involved and requires a calculation. Equation 31 is solved for an S21 or $\frac{P_{rx}}{P_{tx}}$ of -17 dB which was the value observed on average at the center of the band for direct alignment.

$$G_{DUT} + G_{std} - FSPL - Mismatch Loss = \frac{P_{rx}}{P_{tx}} \quad (31)$$

The FSPL 30 inches³⁷ is calculated to be 59.63 dB using equation 4 in section 2.2. The mismatch loss is approximately 0.5 dB across all units tested stemming from a finite S11 of about -10 dB across the band. The gain standard antenna quotes a gain of 24 dBi. Substituting these numbers into the equation above gives:

$$G_{DUT} + 24dBi - 59.63 - 0.5 = -17 \quad (32)$$

$$G_{DUT} = 19.13dBi \quad (33)$$

Therefore, a measured S21 value of -17 dB results gives an antenna gain of approximately 19.13 dBi. This equation is programmed into a spreadsheet and the S21 data obtained is converted into an Gain vs. Frequency curve. To ensure accuracy of the data obtained, the Gain vs. Frequency data was taken at multiple different distances.

This plot, along with the three others mentioned earlier in this section were obtained for each prototype antenna. They are presented in the following sections beginning with the single switch configuration.

³⁷The tx to rx separation distance should be measured from the antenna phase centers which will occur very close to the throat of the horn antennas used here

5.4.2 Single Switch Configuration

This section presents test results for the Single Switch Configuration (AKA 1-way) design with the best performance. The manufactured PCB and horn antenna are shown in Figure 79 and Figure 80 respectively.

Figure 79: 1-way PCB

Figure 80: 1-way horn antenna, PCB and surface mount connector (shown without screws and alignment pins)

The PCB found to perform the best was number 011. It is important to note that this PCB is not identical to the one simulated in section 5.2.1. Figure 81 shows the S_{11} for PCB number 011 in the disabled state.

Figure 81: S11 curve for single switch configuration

A separate unit was made of PCB 011 with the vias shorted to ground to test the enabled state performance. However, the manufacturing process failed on this unit, and so it was not able to be tested. Figure 82 shows the gain vs. frequency plot of the 1-way in the disabled state.

Figure 82: Gain vs. Frequency plot for single switch configuration

The plot shows a gain hovering around 18.5 dBi across a very wide bandwidth of 7 GHz between 28GHz and 35 GHz. Comparing the gain here to the directivity predicted by the simulator (18.5 dBi) gives an efficiency close to 100%! The following uses calculations done in section 5.4.5 to determine whether such a high radiation efficiency agrees with theory.

Substrate and metal loss total 0.145 dB/inch as demonstrated by calculations in section 5.4.5. This value does not include the metal losses associated with the walls of the horn antenna but this will only add 0.043 dB/inch. In addition, connector insertion loss of around 0.1 dB for the coax to PCB connector should be added. This results in total losses around 0.3 dB or 7%. This means that in theory, the antenna should obtain a radiation efficiency of 93%. Therefore, the observed radiation efficiency of around 100% is only different from theory by tenths of a dB which is consistent with test setup error.

The next metric tested is the antenna's cross polarization isolation. Figure 83 shows a cross pol isolation of about 20 dB across the band. This metric is important if dual polarization systems are to be integrated. The 20 dB value provides a good base for such applications.

Figure 83: Cross polarization isolation for directly aligned transmit and receive antennas with tx rotated 90 ° - single switch configuration

Finally, Figure 84 shows the radiation pattern shape in the elevation and azimuth cuts for a frequency of 30 GHz. The shape is almost identical to that predicted by simulation shown in Figures 63 and 64. The sidelobes in the elevation cut are even lower in testing (-13 dB) than what was simulated (-10 dB to -11 dB). In addition, the simulated and tested beam-widths are very similar, with values around 20° in both cuts. From these metrics one can calculate the aperture efficiency using equation 34 to be between 70% and 75%, which is again inline with what was targeted.

$$G = \frac{4\pi A_{phys} e_a}{\lambda^2} \quad (34)$$

Figure 84: 30 GHz radiation pattern for single switch configuration - see Appendix A for more radiation pattern data

From this data it is evident that simulation and testing matched up almost perfectly. The rest of the radiation pattern data for this unit may be found in appendix A. The next section presents very similar results for the 2-way antenna.

5.4.3 Double Switch Configuration

This section presents test results for the Double Switch Configuration (AKA 2-way) design with the best performance. The manufactured PCB and horn antenna are shown in Figure 85 and Figure 86 respectively.

Figure 85: 2-way PCB

Figure 86: 2-way horn antenna, PCB and surface mount connector (shown without screws and alignment pins)

The PCB found to perform the best in this series was number 018. Figure 87 below shows the S11 curve for PCB 018.

Figure 87: S11 curve for dual switch configuration

The S11 is around 10 dB down from 27.5 GHz to 33 GHz. This only means the energy is effectively radiated, but does not say which port the energy escapes out of. The hope of course, is that all the energy is escaping through the disabled port. Aligning the disabled port antenna with the transmit antenna resulted in the gain vs. frequency curve shown in Figure 88.

Figure 88: Gain vs. Frequency plot for dual switch configuration

The gain is around 18.5 dBi to 19 dBi across this same frequency range which means nearly all the power is escaping out of the disabled port. A radiation efficiency between 90% and 100% is again inferred by comparing the gain and directivity plots. Another interesting thing is that simulation results for the 2-way do not agree well with test results. The real antenna outperforms the simulation model with a gain that remains fairly constant over about a 6 GHz bandwidth compared to about 4 GHz in simulation. To determine if the minor variations applied from unit to unit during the PCB run were responsible, unit 018 was modeled in the simulator. The results which are shown in Appendix A were similar to that shown in earlier 2-way simulation data.

Figure 89 below shows a cross polarization isolation of greater than 20 dB for the 2-way. This is a good level of cross pol. isolation and again, is a good starting point for an antenna system that might eventually incorporate a dual polarization scheme.

Figure 89: Cross polarization isolation for directly aligned transmit and receive antennas - dual switch configuration

Figure 90 shows the elevation and azimuth radiation pattern at 29 GHz for the 2-way antenna. Notice a beamwidth of about 20° in both planes which again will result in an aperture efficiency of between 70% and 75%. More radiation pattern data for different frequencies is shown in Appendix A, however much of the data resembles Figure 90.

Figure 90: 29 GHz radiation pattern for dual switch configuration - see Appendix A for more radiation pattern data

The 2-way produces test data that outperforms what was initially anticipated by simulation. In addition, this 2-way design has the opportunity to be made wider band by using techniques described in Chapter 4. Either way, this unit operates very effectively may be used as a building block in larger antenna designs ³⁸. The coming section examines another potential building block called a quadruple switch configuration or 4-way.

5.4.4 Quadruple Switch Configuration

This section presents test results for the Quadruple Switch Configuration (AKA 4-way) design with the best performance. The manufactured PCB and horn antenna are shown in Figure 91 and Figure 92 respectively.

³⁸Such as the 16-way

Figure 91: 4-way PCB

Figure 92: 4-way horn antenna, PCB and surface mount connector (shown without screws and alignment pins)

Figure 93 below shows the S11 curve for PCB 033 which produces the best results. The plot shows an 8dB match between 28 GHz and 32 GHz. Figure 94 shows the gain vs frequency curve produced when the disabled port is aligned with the transmit antenna. The plot displays a gain hovering around 18 dBi to 19.5 dBi which confirms that most of the power seen escaping in the S11 curve gets out through the disabled port. The peak gain is higher because antenna loss is less due to the smaller the distances. Compared to the 2-way where the signal distance is 1.5 inches in the dielectric, the signal inside 4-way must only travel 0.7 inches. This results in a drop in loss between 0.15 dB and 0.2 dB. This, in addition to measurement environment error of 0.1 dB to 0.3 dB, is likely responsible for the discrepancy.

Figure 93: S11 curve for quadruple switch configuration

Figure 94: Gain vs. Frequency plot for quadruple switch configuration

The cross polarization plot of the 4-way is shown in Figure 95. The cross pol. isolation has improved 5 dB and floats around -25 dB. This points to a positive trend relating number of output antennas to cross pol isolation. Cross pol improvements will likely improve significantly for 8-way and 16-way directional antennas.

Figure 95: Cross polarization isolation for directly aligned transmit and receive antennas - quadruple switch configuration

Finally, Figure 96 presents the elevation and azimuth radiation patterns at 29 GHz for this PCB. The plot shows beamwidths of around 20° in both cuts. Again, this would result in an aperture efficiency of around 70%. Another feature present is the side lobes occurring in the elevation cut at -12 dB and -13.5 dB which agrees with simulation.

Figure 96: 29 GHz radiation pattern for quadruple switch configuration - see Appendix A for more radiation pattern data

The next section quantifies metal and substrate loss theoretically and later confirms the calculated figures using test data.

5.4.5 Predicted vs. Measured Substrate & Metal Loss

There are three main kinds of loss in transmission lines; they are dielectric loss, conductor loss, and radiative loss. In general, waveguide structures do not have any radiation loss since all the EM energy is confined within the structure itself. In the waveguide structure being dealt with for this design, there are three places where radiative losses could occur: at the connector to pcb junction, at the pcb to horn junction, and between the pads and the ground plane at each post. Upon further investigation through simulation, loss at these three junctions is determined to be negligible, leaving dielectric loss and conductor loss dominant. The objective of this section is to assign a value to these two losses. Dielectric loss is first to be investigated below.

First, we calculate the wave number 'k',

$$k = \omega\sqrt{\mu\epsilon} = 2\pi(30 * 10^9)\sqrt{\mu_0\epsilon_0 * 2.2} = 932.58 \text{ rad/m} \quad (35)$$

Where ω is the operating frequency (radians), μ is the absolute magnetic permeability of the material, and ϵ is the absolute permittivity of the material. The cutoff wave number ' k_c ' is given as:

$$k_c = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{n\pi}{b}\right)^2} \quad (36)$$

Recall that our transmission mode is TE_{10} , where $m = 1$ and $n = 0$ which simplifies equation 36 to:

$$k_c = \frac{\pi}{a} \quad (37)$$

For a width of 4.4 mm (173 mil), k_c is calculated as:

$$k_c = \frac{\pi}{4.4\text{mm}} = 714 \text{ rad/m} \quad (38)$$

Using the wave numbers calculated above we can calculate the phase constant ' β ':

$$\beta = \sqrt{K^2 - K_c^2} \quad (39)$$

$$\beta = \sqrt{932.58^2 - 714^2} = 600 \text{ rad/m} \quad (40)$$

Finally, the loss of the dielectric, α_d , can be calculated for Rogers 5880 material as follows:

$$\alpha_d = \frac{k^2 \tan \delta}{2\beta} \quad (41)$$

Using a loss tangent for Rogers 5880 of $\tan\delta = 0.0009$, the equation becomes:

$$\alpha_d = \frac{(932.58)^2(0.0009)}{2 * 600} \quad (42)$$

$$\alpha_d = 0.652 \text{ Np/m} = 5.67 \text{ dB/m or } 0.145 \text{ dB/inch} \quad (43)$$

In all three variations manufactured, the signal must travel anywhere from 0.7 inch to 1.5 inches which results in maximum dielectric loss of about 0.22 dB.

The other major loss component is conductor loss. The final product is made with a aluminum which has a conductivity of $\sigma = 3.77 \times 10^7 \text{ S/m}$. The first step is to calculate a value for the sheet resistance, R_s of aluminium at 30 GHz:

$$R_s = \sqrt{\frac{\omega\mu_0}{2\sigma}} \quad (44)$$

$$R_s = \sqrt{\frac{2\pi * (30 * 10^9) * (4\pi * 10^{-7})}{2 * (3.77 * 10^7 \text{ S/m})}} \quad (45)$$

$$R_s = 0.056\Omega \quad (46)$$

Using equation 48 from Pozar cite58, we obtain the loss of conductor, α_c :

$$\alpha_c = \frac{R_s}{a^3 b \beta k \eta} (2b\pi^2 + a^3 * k^2) \quad (47)$$

Where η is the intrinsic impedance of the material filling the waveguide.

$$\eta = \frac{\eta_0}{\sqrt{\epsilon}} = \frac{377 \text{ ohm}}{\sqrt{2.2}} \quad (48)$$

Substituting all variables into Equation 46 gives a conductor loss of:

$$\alpha_c = 0.192Np/m = 1.67dB/m \text{ or } 0.043 \text{ dB/inch} \quad (49)$$

Adding both the dielectric and conductor loss results in a total loss 0.188 dB/inch. For a maximum signal distance of 1.5 inches, the total loss should be 0.2845 dB. In other words, at a max distance of 1.5 inches the loss will only be 6% of the input power. This is a very small loss for such a high frequency of operation.

The loss can also be predicted at 60 GHz for a half sized waveguide using the equation set above. At 60 GHz $\alpha_c = 4.66dB/m$ and $\alpha_d = 11.26dB/m$, resulting in a total loss of 15.92dB/meter (0.388dB/inch). This is roughly twice the loss at 30 GHz.

Figure 97: Dielectric and metal losses for a 1 inch SIW section

Figure 97 shows the total measured dielectric and metal loss for the prototype. The value of the loss shown in S21 is only of interest at intervals where the return loss, S11, is at it's lowest, in a 'null'. Very little reflections occur at these nulls, making the loss measurements more accurate. As you can see, the value of S21 at these nulls is around -0.2 dB. This is very close to the expected value of

0.188 dB calculated above.

6 CONCLUSIONS AND FUTURE WORK

Due to the immense free space path loss at 60 GHz, a high gain, narrow beam antenna solution is required to establish a reliable communication link. This thesis first attempts to identify and describe the problem. Then, an antenna is designed to solve it. Prototypes at 30 GHz were manufactured to test the basic technology and the results obtained were as follows:

1. Gain of 18 dBi to 19.5 dBi
2. Fractional bandwidths of between 14% and 24%
3. Azimuth beam 20°
4. Elevation beam 20°
5. Aperture Efficiency between 70% and 75%
6. Radiation Efficiency between 90% and 100%

The results are truly spectacular and this technology could very well the antenna solution to WLAN room coverage at 60 GHz.

The advantages of this technology are many. The prototypes built have low loss as evidenced by their high radiation efficiency. The substrate & metal losses calculated and tested to be around 0.2 dB/inch at 30 GHz, however, this number will double to about 0.4 dB/inch at 60 GHz. This loss may be compensated for by making the 60 GHz devices two times shorter³⁹.

Since the technology makes use of horn antennas to radiate energy, high gains and high aperture efficiencies are easily obtainable. In addition, both components of the antenna are easily manufacturable. The horn component can be

³⁹This would result in the device being the same number of wavelengths long and therefore would likely behave the exact same all other things properly scaled.

taken care of by any high precision milling machine. The PCB should be made by a high quality PCB shop which can provide maximum accuracy and small tolerances so as to minimize error.

As with any engineering design there are also disadvantages, however, they seem to be few and far apart in this technology. The first potential disadvantage is that the horn aperture at 30 GHz measures 1.25" by 1", which is fairly large, especially if 16 antennas are required around azimuth. However, at 60 GHz the physical area of the aperture would be quartered resulting in approximate aperture sizing of 0.5 inch by 0.5 inch. This is more inline with what might be required by final product. Another potential disadvantage is the diode loss associated with its 4Ω . This loss was quantified in simulation to be 0.5 dB for a single switch. It turns out that this loss gets smaller as more reflectors are added since it is equivalent to putting resistors in parallel.

The next logical step in the development of a complete antenna solution for 60 GHz Wi-Fi is the design and integration of diodes and the control line structure. Once the control line structure is understood, the designer may move to the design of a complete 16-way layer including the control line structure. Successfully integrating a control line and antenna structure for a 16-way antenna would be a major accomplishment.

The only remaining task would be to integrate an additional two layers and figure out how to switch between them. A structure like this would provide almost complete room coverage at 60 GHz with the exception a small chunk of space directly underneath the access point. The solution for this section may be lower gain and broader beam and should only require one antenna. This would effectively be another layer and therefore the structure responsible for switching between layers should also be able to switch to this antenna.

Another potential piece of future work would involve being able to propagate a dual polarization scheme to double available bandwidth! Assuming the design of a single polarization antenna is completed, this task is easy to accomplish. The dual polarization solution would combine two of the single polarization antennas. The only difference between the two would be the addition of a cross polarizing filter placed as a cap on the horns of one of the antennas. This would allow the achievement of dual polarization, with minimal extra effort.

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A Appendix - Simulation Results for Units tested

A.1 Single Switch Configuration - Unit 011 Simulation

Results

Figure 98: 1-way s-parameters - unit 011

Figure 99: 1-way radiation pattern azimuth - unit 011

Figure 100: 1-way radiation pattern elevation - unit 011

A.2 Double Switch Configuration - Unit 018 Simulation

Results

Figure 101: 2-way s-parameters - unit 018

Figure 102: 2-way radiation pattern azimuth - unit 018

Figure 103: 2-way radiation pattern elevation - unit 018

A.3 Quadruple Switch Configuration - Unit 033 Simulation Results

Figure 104: 4-way s-parameters - unit 033

Figure 105: 4-way radiation pattern azimuth - unit 033

Figure 106: 4-way radiation pattern elevation - unit 033

B Appendix - More Radiation Pattern Data for Units Tested

B.1 Single Switch Configuration - Unit 011 - Radiation Pattern Test Data

Figure 107: 1-way complete radiation pattern azimuth - unit 011

Figure 108: 1-way complete radiation pattern elevation - unit 011

B.2 Double Switch Configuration - Unit 018 - Radiation Pattern Test Data

Figure 109: 2-way complete radiation pattern azimuth - unit 018

Figure 110: 2-way complete radiation pattern elevation - unit 018

B.3 Quadruple Switch Configuration - Unit 033 - Radiation Pattern Test Data

Figure 111: 4-way complete radiation pattern azimuth - unit 033

Figure 112: 4-way complete radiation pattern elevation - unit 033

C Appendix - Another Potential 16-way Antenna Arrangement

Figure 113: Another Potential 16-way Antenna Arrangement