Experimental Validation of a 3D-Printed Millimeter Wave/THz Power Combiner

Jacob C. Giese

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Experimental Validation of a 3D-Printed Millimeter Wave/THz Power Combiner

By

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B.S. Electrical Engineering, University of New Mexico, 2019

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ABSTRACT

The terahertz band (THz) in the electromagnetic spectrum is an untapped resource that possesses unique features for nondestructive methods of material imaging and detection. However, little advancements have been made in regard to a practical long range portable THz device. The solution proposed is an oversized cylindrical waveguide termed Power Combiner, designed to combine 12 azimuthally aligned rectangular waveguides to produce power levels suitable for threat detection at secure distances. This paper explores the physical parameters of the power combiner and the power distribution network and creates the 12 signals that allow for modification at the input of the power combiner. The correlation between optimizing the power distribution network and the power combiner’s radiation characteristics are studied for this novelty case.
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1. Chapter 1: Introduction

1.1 Motivation

Most modern technology relies, to some degree, on transporting and receiving information in the form of radiated power characterized by the electromagnetic spectrum. The electromagnetic spectrum constitutes a wide variety of implementing methods of detection. Detection via electromagnetic radiation involves measuring the interaction between the reflected signal and its target. This process can detect a known material based on its interaction with a corresponding frequency, or understanding a new material by transmitting a wide band signal and measuring the frequencies reflected via emission or absorption lines. Two well documented regions of the spectrum are the microwave and optic regimes. However, between these two lies a sub-range known as the terahertz regime (THz) that holds characteristics that may further develop means of detection in life threatening scenarios [1].

The THz band in the electromagnetic spectrum is an untapped resource that possesses unique features for nondestructive methods of material imaging and detection. THz radiation can be used to excite intermolecular vibrations as well as identify chemical species. Another feature is that THz signals may radiate through nonmetallic and non-polar materials, which can lead to better imaging of materials through cardboard, plastic and concealed dry packaging. This is advantageous compared to its x-ray counterpart [2]. These features of imaging sensitivity can be applied to detect explosive and toxic materials though packaging or if it is underground. Through the method of THz time domain spectroscopy (THz-TDS), both pure and plastic explosives can
be detected due to the materials’ nitrogen salts in their structure with resonant lines of absorption in the frequency region of 0.3–10 THz [3]. However, THz-TDS is usually done at very low power levels via a body scanner. Therefore, there are no practical methods of deliver sufficient THz power to operate such a device from a distance. Little advancements have been made in regard to a practical long range portable THz device. The lossy nature of a THz signal in free space, and the current state of THz power sources that can only generate 20-40 mW, which is ineffective at safe distances, suggest that new techniques are needed. There are devices capable to produce output power in the range of hundreds of watts, but such devices are the size of buildings, and therefore are not portable. Studies have shown that microwave oscillators connected to a single pulser result in constructive signal interference and a summation of power, with a phase tolerance of 25% between the combined sources. This has led to this study and simulation of utilizing methods of power combining to incorporate non-stationary oscillators in the attempt to combine the total radiated power in an over-sized 12 port antenna.

1.2 History of Applicable Power Dividing Technology

For the past century, physicist and engineers alike have dedicated their life’s work to the advancement of microwave systems and technology. Although much work has already been done, many of the microwave systems and components used today were created in the 20th century due the need of communications and radar during World Wars I and II. The study of electrodynamics dates into ancient periods where atmospheric charges were invested to understand lighting, which later shifted to static-electricity theory [4]. During the 19th century James Clerk Maxwell discovered the duality and harmonious relationship of electrical and magnetic forces. At the time these field interactions were well understood independently, as formulated by physics such as Faraday, Ampere and Gauss; however, it was not until 1873 that
Maxwell deduced their codependent existence and proved their propagating behavior solely on a mathematical basis. A decade later, Maxwell’s equations (which were simplified into vector notation by Oliver Heaviside) were experimentally proven by the German physicist Heinrich Hertz. During this time, the prolific Lord Rayleigh made the first notion in utilizing metal tubes, later called waveguides, to direct electromagnetic waves and theorize the effects of scattering at boundaries [5]. He hypothesized the fundamental ideas of guided wave characteristics such as mode propagation and cutoff; these ideas would not be developed until the microwave resurgence of World War II due to the growing demand for lower frequencies due to the lack of reliable microwave sources [6].

Although waveguides had been considered theoretically, the ideas at the time were abandoned and were rediscovered in the 1930’s. During this time two theorists were on the cusp of rediscovering the ideas that Lord Rayleigh had theorized decades ago – Dr. George Southworth of Bell Laboratories and Dr. Wilmer Barrow of Massachusetts Institute of Technology. Both published their findings in 1936 where both analyzed circular waveguide geometries, cutoff frequencies, guide wavelengths, and attenuation. Dr. Southworth’s work received skepticism from Bell Labs but was later accepted and published once they knew MIT was performing similar research. Both credited Lord Rayleigh and the work he did in the past. After the two initial publications, additional research was conducted on rectangular waveguide geometries, and it was soon concluded that their configuration led to simpler analytics and field distributions. It was not until the advent of radar during World War II that would spark significant interest in the field of research regarding microwave devices and the continuation of waveguide analytics. During these times, a collaborative effort by MIT, Telecommunications Research Establishment in England, and McGill University proved to be the epicenter of
microwave research, focusing their work on optimizing radar communications and understanding waveguide discontinuities analytically [5].

1.2.1 Magic Tee

In the midst of the rapid advancement of microwave research during World War II, physicists were able to apply theoretical analysis directly to physical experiments with the aid of new microwave measurement capabilities. Soon after, this led to designs and early models of power dividers created at MIT. Some of these early developments were the T-junction and directional couplers, which are still highly practical microwave devices due to their useful characteristics of power combination and division with simple implementation [7]. Along with the Rat Race coupler, the Magic Waveguide Tee was one of the earliest passive devices to have equal division of an input signal with little to no losses. The Magic Tee’s matching capabilities were simple considerations of characteristic impedance in regard to the signal generator, or by incorporating additional components within the interior of the junction [8]. First formulated at Bell Labs by W. A. Tyrell, the initial designs were coined “hybrid” devices, where a signal would enter a 4-way system and be equally split into two paths and be isolated from the third and fourth port. The most well-known physical configuration of a Magic Tee Waveguide is shaped as the letter T, where the top is a colinear junction of 2 waveguides and the bottom is a series connected guide and considered to be a main line. In his early developments, Tyrell connected a coaxial line to the main line which also influenced the T shape. Additional physical developments were in regard to the electric (E) and magnetic (H) fields. However, Tyrell understood that the E-plane and H-plane configurations only differed by a parallel and series
connection of the main line and these two devices could be superimposed to create a hybrid or Magic tee. The hybrid tee configuration can be seen in Fig. 1.1.

Figure 1.1. Typical schematic of a Magic Tee power divider [12].

It should be noted that the power dividers used in the power divider network (PDN) were not in a letter T configuration, but rather a 4-way output; its design will be further discussed in the following chapter; however, the underlying physics are the same. For the purposes of the introduction, a T shaped Magic Tee will be considered. The physics involved in explaining the analytical theories that underly the Hybrid Tee are very cumbersome and generally ignored. The final proof is easier to understand qualitatively, which is why it is given the nickname Magic Tee. Utilizing the Magic Tee as a power divider, a signal enters the parallel connected port (also labeled as the Sum Port in other publishing’s [8]), and due to the symmetry of the Magic Tee, the equal phase and amplitude are split between the colinear components. Because of the even characteristics of the incoming wave and the odd symmetry of the perpendicular port (series, or \( \Delta \) port) the two do not couple, therefore creating complete isolation in the series port given the ideal case [9]. It can also be seen due to the systems reciprocity that the two signals can enter the device and combine out of one port (the sum port) while isolated from the \( \Delta \) port.
1.2.2 Wilkinson Power Divider

Although this type of power divider was not used in the experiments, it utilizes similar characteristics of matching and isolations to the Magic Tee and, therefore, will also be examined.

It was not until two decades after the end of World War II that microwave researchers began to focus on their attention on passive devices for stripline and waveguide applications [6]. Various publications cite Ernest J. Wilkinson as a pioneer in power dividing theory and techniques. In 1960, he published his work with N-way hybrid power dividers (later termed Wilkinson Power Dividers) [9]. This publication laid the foundation of the modern reciprocal power divider. The most favorable characteristic of the Wilkinson power divider provides isolation between ports, i.e., with S-parameter \( S_{ii} = 0 \). This initial model was the take-off point in matched reciprocal power dividers and is worth mentioning due to the shared characteristics of the power dividers used in the novel PDN. First hypothesized in 1960, Wilkinson’s original power divider experiment was attempted using a coaxial line where a hollow inner conductor was split into \( N \) separate conductors with characteristic impedance \( Z_0 \). Each conductor had length equal to a quarter wavelength \( \lambda/4 \) with a shunt resistor connected radially in-between.

Wilkinson formulated that choosing a value for \( Z_0 = \sqrt{N} \) and setting the value of the resistor \( R = 2*Z_0 \) will result in an entirely matched system due to the nature of the parallel configuration. Using lengths equal to \( \lambda/4 \) ensures that any reflection of the outputs will undergo deconstructive interference. This was also verified for input signal on the output ends and therefore can be used as a power combiner [9].

1.3 Preceding Research

1.3.1 Power Combining Antenna

The predecessor to this experiment was the simulation and design of a power combining
antenna labeled the Power Combiner. It was proven theoretically that a power combining method to produce high power radiation in the THz regime could be accomplished by combining THz signals into a single circular waveguide to excite a TE$_{01}$ output [3]. The paper explored the need for, and the utility for, a portable THz signal radiator. Moreover, the paper interprets the design of the Power Combiner and its characteristics that could be altered to achieve optimal results.

The Power Combiner design, shown in Fig. 1.2, implements multiple rectangular waveguides feeding an oversized coaxial guide that ultimately produces a symmetric cylindrical wave. The 12 rectangular waveguides are azimuthally aligned around the input of the Power Combiner such the system maintains symmetry. Utilizing the dominant TE$_{10}$ mode, the 12 inputs travel between the coaxial channel, and with the help of symmetry, a single TE$_{01}$ mode is synthesized and radiated. If the electric fields are orientated in the radial direction, the output field would be TM$_{01}$ polarized, making the center circular channel radius larger and therefore higher order parasitic modes would be created. With a radial configuration, a maximal efficiency would only reach 92% at a single frequency of 350 GHz. However, when the Power Combiner is fed with azimuthal E-field inputs, the power combiner will produce a TE$_{01}$ mode with a
maximum of 100% efficiency obtained at a single frequency of 500 GHz while having efficiencies of 93% and higher for a frequency band of 0.38-0.5 THz.

*Figure 1.3. Simulated TE$_{01}$ field input.*

*Figure 1.4. Power Combiner’s simulated radiation pattern.*

This efficiency is explained by observing the boundary conditions on tangential E. At the
walls of the Power Combiner, the tangential (azimuthal) components of the electric field must go to zero for a perfect electrical conductor (PEC). At the boundary there is no field, and therefore, no surface current. By negating these losses, high efficiencies are achievable in high frequency regimes. Transitions from a PEC to a realistic copper conductor in the simulations showed about a 10% drop in efficiency, but the efficiency still rose due to the boundary conditions on tangential E and H. The 12 ports must be fed in phase to negate any interference between signals. To test the simulated results, the system operated in a scaled down frequency band that operated in the high GHz regime (chosen to be 75-82 GHz) rather than the band that used in simulation. This was due to the fact that THz devices are often costly and were unpractical for this experiment. The Power Combiner design was then fabricated via 3D printing. The idea of fabricating using a 3D printer, as opposed to a custom order through an RF manufacture, is a noteworthy characteristic because it implies sensitive and high performing technology can be made in a common laboratory setting. Then, a layer of copper plating was applied to all surfaces of the physical Power Combiner. To feed the 12 waveguide inputs, a PDN is utilized as the input to the Power Combiner, along with an extra adapter placed in-between the PDN and the Power Combiner to ensure each port is aligned and matched to the port size of the Power Combiner. This experimental setup is the basis of this research.

1.4 Thesis Organization

The subsequent methods of testing and data acquisition are as follows. Chapter 2 formulates the basic principles that constitute the physics behind the PDN and Power Combiner. Chapter 3 is dedicated to the multiple PDN configuration attempts to properly feed the Power Combiner. Chapter 4 is dedicated to experimental parameters and result analyses. Chapter 5 is posed as a discussion interlude where ideas, shortcomings, solutions, and future works are
proposed. Finally, Chapter 6 will conclude the thesis and summarize the work presented.
2. Chapter 2: Fundamentals and Experimental Setup

As mentioned in the previous chapter, a PDN was the intermediate component used for signal manipulation to achieve a balanced feed for the Power Combiner. The PDN was utilized so that the input signal may be divided into 12 channels, each having passive operational devices so that the signals may be adjusted manually. This gives attenuation and phase control between the 12 channels, as well as incorporating signal isolation devices. The 12 parallel channels are reconfigured such that the waveguide channels match the cylindrical alignment of the input to the Power Combiner. The design is depicted in Fig. 2.1.

![Figure 2.1. Photograph of the E-band power distribution network (PDN).](image)

The PDN that feeds the 12 signals into the power combining antenna was fed by a millimeter waveguide extender connected to a Vector Network Analyzer via a coaxial cable. The millimeter waveguide extender is used to operate at the higher frequencies of the microwave range in the lower THz region. The signal radiated by the millimeter waveguide extender is delivered by an E band rectangular waveguide. Analytically, the behavior of the signal inside the waveguide and PDN can be derived using Maxwell’s equations and the boundary conditions on
the gold-plated interior of the waveguide and PDN. Therefore, before further explanation of the components of the PDN, fundamental background is necessary to understand how the wave will be manipulated. Details of the components will be explored later in the chapter.

2.1 Background Electromagnetic Theory

An electromagnetic signal is injected into the PDN that serves as the input to the system. Fundamentally, this can be further investigated by understanding the initial concepts of Maxwell’s equations and how they are related to provide the general wave solution without context to the specific environment using the equations

\[ \nabla \times \mathbf{E} = - \mathbf{M}_i - \frac{d \mathbf{B}}{dt} \]  
(2.1)

\[ \nabla \times \mathbf{H} = - \mathbf{J}_i + \mathbf{J}_c + \frac{d \mathbf{D}}{dt} \]  
(2.2)

\[ \nabla \cdot \mathbf{D} = - q_{ev} \]  
(2.3)

\[ \nabla \cdot \mathbf{B} = - q_{mv} \]  
(2.4)

where

- \( \mathbf{E} \) = electric field intensity (volts/meter)
- \( \mathbf{H} \) = magnetic field intensity (amperes/meter)
- \( \mathbf{D} = \varepsilon \mathbf{E} \) = electric flux density (coulombs/square meter)
- \( \mathbf{B} = \mu \mathbf{H} \) = magnetic flux density (webers/square meter)
- \( \mathbf{M}_i \) = impressed magnetic current density (volts/square meter)
- \( \mathbf{J}_i \) = impressed electric current density (amperes/square meter)
- \( \mathbf{J}_c \) = conduction electric current density (amperes/square meter)
- \( q_{ev} \) = electric charge density (coulombs/cubic meter)
\( q_{m,v} = \text{magnetic charge density (webers/ cubic meter)}. \)

It should be noted that values \( \mathcal{M}_i \) and \( q_{m,v} \) are fictitious quantities not realizable in the real world. This is due to the dipole property of any given magnetic flux. However, by generalizing the concept of current and current density to allow \( \mathcal{M}_i \) and \( q_{m,v} \) into equations (2.1) and (2.4), a symmetry between \( \mathcal{B} \) and \( \mathcal{D} \) can now be seen across all for equations. Utilizing \( \mathcal{M}_i \) and \( q_{m,v} \) as source quantities greatly reduces complex boundary analytics [10].

Manipulating these equations with vector analysis yield wave solutions that can be specified to rectangular waveguide conditions. For a source free, linear, homogeneous region, Maxwell’s equations may be written in phasor form by:

\[
\nabla \times \vec{E} = -j \omega \mu \vec{H}, \tag{2.5}
\]

\[
\nabla \times \vec{H} = j \omega \varepsilon \vec{E}. \tag{2.6}
\]

This gives two equations and two unknowns. It also allows \( \vec{E} \) and \( \vec{H} \) to be solved in terms of one another by taking the curl of both equations utilizing the vector property

\[
\nabla \times \nabla \times \vec{A} = \nabla(\nabla \cdot \vec{A}) - \nabla^2 \vec{A}. \tag{2.7}
\]

Equation (2.5) becomes:

\[
\nabla \times \nabla \times \vec{E} = j \omega \mu \nabla \times \vec{H} = \omega^2 \mu \varepsilon \vec{E}, \tag{2.8}
\]

which simplifies to

\[
\nabla^2 \vec{E} + \omega^2 \mu \varepsilon \vec{E} = 0 \tag{2.9}
\]

\[
\nabla^2 \vec{H} + \omega^2 \mu \varepsilon \vec{H} = 0, \tag{2.10}
\]

where the propagations constant may be defined as \( \beta = \omega \sqrt{\mu \varepsilon} \). This gives the form of two second order differential wave equations [6]. In the case of rectangular geometries, the solutions to the equations above can be solved by treating the \( \hat{x} \) \( \hat{y} \) \( \hat{z} \) components of the E field individually by the separation of variables method, thus producing the solution
\[ \overline{E}(z) = E_0 e^{-j(\omega t + \beta z)} + E_0 e^{j(\omega t + \beta z)} \] (2.11)

\[ \overline{E}(x) = E_0 \cos(\omega t + \beta x) + E_0 \sin(wt + \beta x) \] (2.12)

\[ \overline{E}(y) = E_0 \cos(wt + \beta y) + E_0 \sin(wt + \beta y). \] (2.13)

These solutions assume the wave is traveling with phase progression in the \( \hat{z} \) direction. The \( \hat{x} \) \( \hat{y} \) components of the \( \overline{E} \) field represent standing waves that occur for the non-transverse components of the wave. Each component of the \( \overline{E} \) field is considered to be lossless. In practical applications it is often required to include losses obtained from the system. However, for the purposes of feeding the power combining antenna, characterizing the phase progression of the PDN is more substantial than accounting for loss. Losses found in the system may be measured and reported but phase balancing requires greater attention [10].

2.1.1 Boundary Conditions

The three classical examples used to describe electromagnetic waves at an interface are: general, lossy, and perfect electrical conducting boundaries (PEC) [6]. The derivation for the general and dielectric conditions are not necessary but can be derived using Maxwell’s equations in integral form at an arbitrary boundary for tangential and perpendicular situations [25]. The equations are

\[ \hat{n} \cdot \overline{D}_1 = \hat{n} \cdot \overline{D}_2 \] (2.14)

\[ \hat{n} \cdot \overline{B}_1 = \hat{n} \cdot \overline{B}_2 \] (2.15)

\[ \hat{n} \times \overline{E}_1 = \hat{n} \times \overline{E}_2 \] (2.16)

\[ \hat{n} \times H_1 = \hat{n} \times H_2. \] (2.17)

However, the copper and gold exteriors of the PDN and the Power Combiner follow closely to the PEC condition and was relevant to the experiment and calibration methods. PEC boundary conditions can be explained with a variant of equations (2.14-2.17) as shown below:
\[ \hat{n} \cdot \vec{D}_1 = \rho_s \] (2.18)
\[ \hat{n} \cdot \vec{B} = 0 \] (2.19)
\[ \hat{n} \times \vec{E} = 0 \] (2.20)
\[ \hat{n} \times \vec{H} = \vec{j}_s, \] (2.21)

where (2.14-2.15) pertain to the perpendicular components and (2.16-2.17) pertain to the tangential components, respectively.

The variation between (2.14-2.17) and (2.18-2.21) is due to the infinite conductivity at the boundary. This suggests all free electrons exist at the surface of the boundary, which can lead to three results. First, no fields can exist within the boundary \( E_2 = D_2 = B_2 = 0 \). Second, since all the charge is at the surface, there exists a surface charge density and, therefore, an electric flux density as seen in equation (2.18) and through these charges, a current is induced given a tangential magnetic field in equation (2.21).

### 2.1.3 Microwaves in a Rectangular Waveguide

![Figure 2.2. Schematic of a rectangular waveguide [6].](image)

Equations (2.11) (2.12) (2.13) given above are the general cases of wave propagation. To further develop the physics of the PDN, these solutions need to be considered for rectangular
waveguide boundaries. For the remainder of the derivation only the traveling \( z \) component will be considered and the \( e^{-j\omega t} \) will be omitted. Given that a rectangular waveguide is three dimensional, the wave solutions can be simplified further to solve for the transverse fields. In regard to the PDN, the input signal is transverse electric, meaning that the input signal does not have an \( \vec{E} \) component in the \( \hat{z} \) direction. In order to satisfy the wave equation within a rectangular waveguide for a TE wave, the following Helmholtz equation must be satisfied:

\[
\left( \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + k_c^2 \right) H_z(x, y)e^{-j\beta z} = 0, \quad (2.18)
\]

where the derivative with respect to \( \hat{z} \) is omitted because here the focus is on the \( \hat{z} \) component of the magnetic field as a function of \( \hat{x}, \hat{y} \). This can be differentiated using a separation of variables technique. The solution for the TE_{10} mode in a rectangular waveguide is given by

\[
H_z = A_{10} \cos\left( \frac{n\pi x}{a} \right) e^{-j\beta z} \quad (2.19)
\]

\[
E_y = -\frac{j\omega \mu_a}{\pi} A_{10} \sin\left( \frac{n\pi x}{a} \right) e^{-j\beta z} \quad (2.20)
\]

\[
H_x = -\frac{j\beta \alpha}{\pi} A_{10} \sin\left( \frac{n\pi x}{a} \right) e^{-j\beta z} \text{ where } n = 1, 3, 5, ... \quad (2.21)
\]

\[
E_x = E_z = H_y = 0, \quad (2.22)
\]

where

\[
k_c = \sqrt{k^2 + \beta^2}, \quad (2.23)
\]

\( k \) is the propagation constant of the material inside the waveguide and \( \beta \) is the propagation constant of the signal flowing in the material not within the waveguide. Often this equation can be expressed in terms of \( \beta \) to show that \( \beta \) is generally a function of frequency and the geometric properties of the guide. It is important to note that this is the final form of the solution set. The processes in-between solving the Helmholtz and the final solution involve utilizing the transverse
field solutions in free space and converting the H fields to E fields so that boundary conditions may be satisfied. The coefficients in front are the result of the conversion process between E and H. However, the sinusoidal functions and their arguments are a result of the boundary conditions on tangential E. The electric field cannot exist at the left and right, upper and lower walls of the waveguide due to the boundary condition on the tangential components of the electric field. The boundary conditions were satisfied by implementing the sinusoidal argument \( \left( \frac{n \pi x}{a} \right) \). Additional sinusoidal terms exist in the general solution for a rectangular waveguide, but were cancelled out because their arguments contained the modal parameter n (e.g. TE_{mn}) in the numerator of their argument, which is zero for the TE_{10} case. The frequency band used during testing was a 7 GHz band between 75-82 GHz. It should be noted that this frequency band was not considered to be in the THz regime; the reason for this will be explained in later chapters. The cutoff (lowest) frequency for the dominate mode (TE_{10}) inside a W-12 waveguide is approximately 48.35 GHz [6]. Although there is almost 30 GHz between the cutoff and operating frequencies, the system is still operating in the TE_{10}. This can be seen by using the following equation that accounts for modal parameters based on the propagation constant

\[
f_{c_{mn}} = \frac{1}{2\pi \sqrt{\mu \varepsilon}} \sqrt{\left( \frac{m \pi}{a} \right)^2 + \left( \frac{n \pi}{b} \right)^2}.
\]  

(2.24)

The second propagating mode occurs at approximately 96 GHz.

This is the qualitative basis of how the signal can be modeled in the PDN. Every component of the PDN is based around a rectangular waveguide configuration. The operable components of the PDN are the attenuators and phase shifters, which in their most open state (adjusted to have no effect on the signal) are, in principle, rectangular waveguides.
2.2 Components of the PDN

2.2.1 Power Divider

As the initial signal enters the input port it undergoes a 4-way split via the first power divider, as shown in Fig. 2.3. The 4 new signals each enter another 4-way power divider. Attached to an output port on each power divider in the second set is a 50 Ω matched load, which then creates the 12 signals that will progress through the PDN. Due to the power divider’s necessity to create the equal phase and amplitude of 12 channels, it is worth mentioning its attributes and characteristics.

According to the data sheets of the PDN, the power divider configuration was chosen in regard to well isolated ports, a requirement for power combing and/or splitting [11]. (The two power dividers mentioned in Chapter 1 would both be suitable for the PDN purpose). Due to the frequency of operation, it was designed by Evarant to use a 4-output Magic Tee with a parallel output configuration with internal rods that assist in matching [12]. As seen in Fig. 2.3, the 4-way power divider employs the same principle of symmetry as the Magic Tee. Two pins are
located on each side of the divider with a third pin near the input. The ports of the power dividers are matched and isolated to about -15 dB from neighboring ports and -25 dB from non-neighboring ports. Because no power dividers can be matched, isolated, and lossless, this implies a mismatch at the input port. This can be seen in the datasheet describing an insertion loss of 0.5 dB [13]. The typical insertion loss between the input port and each output port were measured to be -7 dB with plus or minus 0.4 dB of error. To verify the datasheet and ensure equal phase and amplitude at each port, all five dividers were measured using the VNA. The ports not being measured were shunted with a matched load.

2.2.2 Attenuators

The phase shifters and attenuators act as adjustable corrective means to ensure that the 12 signals arrive to the antenna in an optimal configuration to allow maximum radiation. Each arm of the PDN has an attenuator and phase shifter pair connected in series. Both attenuators and phase shifters are controlled and adjusted with a mechanical micrometer that offers a continuous fine tuning to allow repeatable values.

For the purposes of achieving optimal input into the power combining antenna, 12 signals of equal amplitude should be emitted into the antenna. The attenuators were adjusted to account for signal loss as they made their way through the PDN and adjust such that each signal is equal in amplitude. The amplitude of the signal is not of the upmost concern, for as long as the signal is not significantly attenuated, it can be accounted for. Rather, a greater concern is phase balancing, and that process will be discussed in the following section. However, the dielectric inside the attenuators was shown to have complex permittivity and therefore altered the phase of the signal as well. To some degree, this was utilized to aid in phase balancing. Each attenuator is an E-band level setting attenuator [15].
The micrometer operates on a millimeter scale, and tightening the micrometer lowers a lossy dielectric disk into the attenuator’s waveguide cavity. Analytically, two scenarios of the attenuator can be investigated by utilizing boundary conditions mentioned in section 2.1. The first scenario is when the micrometer is fully inserted such that the lossy dielectric is fully across the center of the waveguide. The second scenario is when the micrometer is arbitrarily fastened. If the micrometer is fully loosened the lossy dielectric will have no contact with the signal and therefore the attenuator will act as a rectangular waveguide.

![Attenuator diagram when the dielectric disk is fully inserted.](image)

*Figure 2.5. Attenuator diagram when the dielectric disk is fully inserted.*
A derivation of the first scenario is more intuitive, and the idea can be expressed qualitatively for the second scenario. Assuming a TE$_{10}$ wave entering the attenuator, the fields in the waveguide can be modeled by using Helmholtz equation

\[
\left( \frac{\partial^2}{\partial x^2} + k_{cd}^2 \right) h_z = 0 \quad (2.25)
\]

\[
\left( \frac{\partial^2}{\partial x^2} + k_{ca}^2 \right) h_z = 0, \quad (2.26)
\]

where the partial derivative with respect to $\hat{y}$ is zero because there is no change in the $\hat{y}$ direction [6]. The value $k_d$ is the cutoff propagation constant inside the dielectric and $k_{ca}$ is the air on each side of the dielectric disk given by

\[
k_{cd} = \sqrt{k_d^2 + \beta_d^2} \quad (2.27)
\]

\[
k_{ca} = \sqrt{k_a^2 + \beta_a^2}. \quad (2.28)
\]

These cutoff wavenumbers are different due to the dielectric, but the propagation constants $\beta_d$ and $\beta_a$ are equal due to the phase matching condition given by

\[
\beta_d = \beta_a \Rightarrow \sqrt{\epsilon k_0^2 - k_d^2} = \sqrt{k_0^2 - k_a^2}, \quad (2.29)
\]

where $\epsilon = j\omega \sqrt{\mu \epsilon'}(1 - j\tan\delta)$ which represents the loss in the dielectric of the attenuator and the imaginary component represents the phase change. Equations (2.25-2.26) are solved by the use of separation of variables and $H_z$ and can be converted to $E_z$ using Maxwell’s equations. The solution set to the electric field will be given in three parts: one for each side of the dielectric and one inside the dielectric. Omitting the $H_z$ and $E_z$ the fields inside the attenuator are given by
\[ E_y = \begin{cases} \frac{j\omega \mu_0}{k_a} [A \sin k_{ca}(x) - B \cos k_{ca}(x)] & \text{for } 0 \leq x \leq t_1 \quad (2.30) \\ \frac{j\omega \mu_0}{k_d} [-C \sin k_{cd}(x) + D \cos k_{cd}(x)] & \text{for } t_1 \leq x \leq t_2 \quad (2.31) \\ \frac{j\omega \mu_0}{k_a} [E \sin k_{ca}(a-x) - F \cos k_{ca}(a-x)] & \text{for } t_2 \leq x \leq a, \quad (2.32) \end{cases} \]

where the constants A-F are general constants that account for the reflection and transmission coefficients and are found using boundary conditions at the walls of the attenuator and the two sides of the dielectric. At \( x = 0 \) and \( x = a \) the fields go to 0; therefore, B and F must be equal to 0. The boundary conditions at \( x = t_1 \) or \( t_2 \) state that the fields on both sides of the boundary must be equal. This implies symmetry and the solution of one side may be used for the other. Solving for \( \beta \) numerically shows the proper phase matching condition. The attenuation in the solution is in regard to \( k_d \) however, \( E_y \) is only the transverse component. Additional losses can be found in the direction of propagation. The total E field is

\[ \overline{E} = E_z e^{-j\gamma d}, \quad (2.33) \]

where

\[ \gamma = \alpha + j\beta, \quad (3.34) \]

\( \alpha \) is the conduction loss, and d is the thickness of the dielectric. This solution was formulated by using the wave solutions for a rectangular geometry and the particular solution can be found once the boundary conditions were satisfied. This approach can be used further to find a solution set to the second case where the micrometer is arbitrarily fastened/loosened. It would differ from the solutions above due to the new dielectric change in the \( \hat{y} \) direction and therefore having a non-zero \( h_y \) Helmholtz solution. However, the general process of obtaining the solution is still equivalent.
2.2.3 Phase Shifters

![Phase Shifter Image]

*Figure 2.6. Photograph of the E-band micrometer-driven phase shifter.*

The phase shifters used in the PDN (Fig. 2.6) operate in a similar fashion to the attenuators mentioned above. They are adjustable via a micrometer knob akin to the attenuators and their underlying physics are derived with the same procedure. The difference between the two devices is the dielectric material that interacts with the signal. The phase shifter utilizes a low loss dielectric material where $\alpha$ is essentially zero. This insinuates any micrometer adjustment that alters the boundary condition further is only in regard to $\beta$.

2.2.4 Faraday Isolators

![Faraday Isolator Image]

*Figure 2.7. Photograph of the E-Band Faraday isolator.*
The attenuators and the phase shifters were reciprocal devices due to their design. Because of this, however, reflections from the output could directly interfere with modifying the signal. Therefore, an isolator is required between the phase shifter and the PDN adapter. The best suited isolation configuration would be a Faraday rotation isolator that utilizes ferrimagnetic materials to ensure high nominal isolation, loss insertions, and high bandwidth.

Faraday isolators are characterized as having good transmission and high isolation due to the physics of Faraday rotation. The atoms in materials like iron are highly magnetic polarizers that are very susceptible to magnetic fields which excites electrons that causes orbital movement around the positively charged nucleus [14]. Including a static magnetic field, in turn, creates a torque on the electron. The torque is further described using quantum mechanics but macroscopically can be accounted for in the constitutive parameter $\mu$ becoming a tensor value.

When a field is present on the material, a torque is created that causes a precession of electrons in either clockwise or counterclockwise directions, based on the polarization of the wave (LHCP or RHCP). Therefore, if a static field in the $\hat{z}$ direction is applied to the ferrite and a linear polarized wave is introduced, the interactions produce a tilt angle that is dependent on the length of the ferrite, and the direction of travel. In addition, the tilt angle rotation is in the same direction, even if the direction of propagation is reversed, making the ferrite non-reciprocal.

Using this feature, an isolator can incorporate two resistive slides that are orientated at an angle perpendicular to the forward going wave, and parallel to the backward traveling wave. This is accomplished by utilizing the tilt angle shift. If the tilt angle shift in one direction is 45°, then a reflected wave is shifted by 90° the required perpendicular to parallel polarization shift is accomplished.
2.2.5 PDN Adapter and Flange

The two final components of the PDN are the adapter and flanged guides that create the waveguide configuration necessary to fit onto the input of the power combing antenna. First, the adapter receives all 12 signals that were originally aligned parallel and realigns the guides to be azimuthally aligned, as seen in Fig. 2.8.

![Figure 2.8. Photograph of the custom manufactured PDN waveguide adapter.](image)

The bends and turns in the adapter do not contribute any discontinuities and the varying lengths between each channel are supposed to differ by a multiple of a wavelength [16]. Finally, the flanged component is the connecting piece between the PDN and the antenna. The 12 signals are tapered toward the center to align to the inputs of the antenna.

![Figure 2.9. Photograph of a) Power Combiner adapter front side; b) back side.](image)
3. **Chapter 3: PDN Calibration Process**

3.1 **Considerations of Lossy Components**

The purpose for the calibration process is to account for the real and lossy characteristics of the PDN and Power Combiner. The behavior of some components showed more variation as described in the datasheets from a 75-82 GHz frequency band. The reason the system operated in E-band was due to the fact that THz devices are often costly and were unpractical for this experiment. Therefore, the experiment was scaled down to this frequency regime. To verify the datasheet and to ensure equal phase and amplitude deviation at each port all five power dividers were measured via $S_{21}$ readings while matched loads were attached to the non-tested ports. The majority of the power dividers showed agreement with the specifications listed in the datasheet. The highest measured variation was approximately 13° of phase difference between ports with a maximum of 1 dB difference in attenuation between two ports. The minor phase differences can be accounted for via the phase shifter.

With regard to the attenuators, the attenuation range was 30 dB, which was the value listed on the datasheet. However, changes in the attenuator did not behave linearly based on the micrometer adjustments; rather, different positions of the micrometer accounted for different incremental changes. As mentioned earlier, the attenuators were designed with a dielectric that
also alters the signals phase. When further characterizing each attenuator, it could be seen that the changes in the phase were just as significant as the changes in attenuation.

![Figure 3.1. Attenuator’s attenuation and phase progression.](image)

As seen in the in Fig. 3.1, the phase range can be upwards to 140°; this phase change did prove to behave fairly linearly. Toward the tighter end of the micrometer, the attenuation and phase became volatile, and did not settle until the micrometer was adjusted closer to their center values. At the loosest value the phase and attenuation had little change.

The phase shifters and isolators behave similar to what was stated in the datasheets. The phase shifters were listed as having a 180° phase band, and it was apparent that after each measurement the phase shifters’ lowest band was 190° while the largest was 240°.
3.2 PDN Calibration Methods

Initially it was assumed that the PDN was created in a “balanced” state, where the 12 signals arrived at the end of the PDN adapter in co-phase. After discovering that was not the case, it was presumed that only minor adjustments would be needed to be made. However, initial testing of the PDN without altering any of the physical characteristics showed the input signals’ phases were unbalanced. Therefore, much consideration was given to calibrating the adjustable components of the PDN such that the power combiner would operate with ideal input, thus providing the best results during experiments. The original goal of the calibration process was to adjust the characteristics of the PDN such that both attenuation and phase were matched at the input of the power combiner. However, with the added complexity of the phase changing attenuators, it was decided that only the phase needed to be matched at the power combiner input. For the Power Combiner’s analysis, any attenuation needed to be accounted for. However, this was not the case in regard to the phase; an unmatched phase configuration could lead to destructive interference, thus leading to inaccurate data which would reflect poorly on the Power Combiner [17]. The calibration process proved challenging due to the frequency of operation, specifically when operating within the band of 75-82 GHz, which translates to a wavelength on the order of millimeters. For a 12 port Power Combiner to have good efficiency, it was necessary to try and have as little phase tolerance as possible. It was decided to have a tolerance of no more than 10 degrees. This implies that any physical differences in channels greater than 0.001 m is enough to shift the wave by 10° [18]. This required precise measurements to ensure that each test was congruent with the last. Another layer of complexity came via the custom design of the power adapter piece that is responsible for aligning the 12 waveguides with the Power Combiner. As seen in Fig 2.8 the output port of the adapter was not flanged and, therefore, cannot be
connected to a VNA. A multitude of different techniques were utilized in hopes to balance the phase and to do so with repeatable and accurate methods.

3.2.2 Full Setup

Early attempts to balance the PDN to properly feed the Power Combiner relied on measuring each port and adjusting the phase shifter and attenuator accordingly. The input signal into the PDN was fed from a Vector Network Analyzer, as shown in Fig. 3.2.

![Photo](image)

*Figure 3.2. Photograph of the initial testing with the Power Combiner.*

In the initial stages of testing, a signal traversed through the PDN until it reached the end of the PDN’s adapter. There, a shunt waveguide probe connected to the receiving (Rx) frequency extender was pressed against the output end of the adapter to complete the two-port system to analyze the $S_{21}$ characteristics. The millimeter waveguide extender’s height had to be continually adjusted to account for the azimuthally aligned ports of the adapter. Attached to the input waveguide extender was a rotary waveguide that is capable of 360° rotation so that the input may be aligned to the PDN adapter, as shown in Fig. 3.3.
Each channel was measured separately with the VNA while the other 11 ports of the adapter were shorted with copper tape as to not allow any unwanted signal interaction, as can be seen in Fig. 3.4.

Figure 3.3. Photograph of the rotary waveguide used to align the input.

Figure 3.4. Photograph of the initial testing setup to find the network calibration.
The ports of the adapter varied in height and the size constraints of the millimeter waveguide extender and the PDN adapter led to the only viable method of elevation which was to utilize paper stacks to adjust for the heights. To begin the calibration process, an arbitrary value was chosen to be the standard such that each channel’s $S_{21}$ phase and attenuation would equal this value. After multiple attempts, it was seen that there was no combination of the attenuator and phase shifter that would allow for a complete attenuation and phase balance. Namely, the phase was the issue, even with the phase shifters full band ranges. The first solution was to characterize each attenuator and phase shifter by recording the full band micrometer readings of each device to deeply understand both components’ behavior. This led to the discovery of the attenuators’ phase dependence (mentioned earlier) and the full phase band from each phase shifter. Although each phase shifter had a 200º band, the range of degrees covered by each varied. During the phase matching process, it was seen that different phase values set to be the standard would give different results. The best phase values could match 10 out of the 12 ports and these ports varied as the phase value changed. Understanding the range (band) of degrees each phase shifter covered, certain phase shifters were swapped between channels in hopes to find an optimal configuration. However, no matter what phase point was chosen, there were always two channels that could not be match given the range of the phase shifters. Although a balanced output could not be configured, it was not the fault of the attenuators and phase shifters; rather, it was concluded that this experimental setup proved to be inaccurate due to the insecure connection between the shunt waveguide and adapter. This was proven by the various amplitude and phase variations between tests. It was seen that after the shunt waveguide was set, any small perturbances could completely alter the aligned state of the shunt waveguide to the PDN adapter, which violated the phase tolerance mentioned prior. The instability of the
stacked papers furthered the inaccuracies, as well as the difference in port sizes between the adapter and waveguide (i.e. WR-2 and WR-12) [3,19]. Here it was determined that all parts must be fastened with screws and aligned with prongs.

Ruling this method as ineffective proved that the configuration of the PDN needed to be done with additional care and each reading required a secure connection to the VNA. This proved to be difficult when including the PDN adapter. A full phase range of the entire device, excluding the PDN adapter, could easily be measured and if a proper characterization of the PDN adapter was recorded the solution to a balanced PDN could be achieved. Therefore, the following work was dedicated to further analyzing the PDN adapter.

3.2.3 Focus on the Waveguide Adapter

3.2.3.1 Utilizing Nicolson Ross Weir Method

Understanding the PDN’s adapter was only flanged on one side, the first method to characterize the adapter was to short the non-flanged end while measuring the \( S_{11} \) parameters from the flanged end. Then, the \( S_{11} \) readings could be converted into \( S_{21} \) values using intermediate steps found in the Nicolson-Ross-Weir (NRW) method. The NRW method is a formulative technique to solve for an unknown value of complex permittivity and permeability. This does not align with the purpose for characterizing the adapter. However, in the paper by Rothwell et al., they discuss such a technique to convert \( S_{11} \) to \( S_{21} \) [20]. At the time of the experiment, it was assumed that the DUI was reciprocal, lossless, and symmetric. Applying these conditions to power dissipation in the system, the S-parameters followed unitary properties. Therefore, if the two ports were assumed to be reciprocal, lossless, and symmetric, the follow S-parameter matrix properties apply:

\[
S_{11} = S_{22} \quad (3.1)
\]
\[ S_{21} = S_{12} \quad (3.2) \]

\[ |S_{11}|^2 + |S_{21}|^2 = 1 \quad (3.3) \]

\[ |S_{21}|^2 + |S_{22}|^2 = 1 \quad (3.4) \]

\[ S_{11}S_{12}^{*} + S_{21}S_{22}^{*} = 0. \quad (3.5) \]

Equations (3.2-3.5) lead to the following condition given that the phase of the \( S_{ij} \) is represented as \( \varphi_{ij} \)

\[ \varphi_{11} - \varphi_{12} - \varphi_{21} + \varphi_{22} \pm \pi = 0 \quad (3.6) \]

if the adapter channels are also reciprocal such that \( S_{12} = S_{21} \). This equation simplifies to

\[ \varphi_{11} - \varphi_{21} \pm \frac{\pi}{2} = 0. \quad (3.7) \]

From here, a solution to \( S_{21} \) takes the form

\[ S_{21} = j(\pm 1)S_{11} \sqrt{1 - \frac{|S_{11}|^2}{|S_{11}|^2}}. \quad (3.8) \]

To implement this equation, it was first tested on three different waveguides, each with bends similar in shape to the adapter. This was to obtain accurate \( S \)-parameter measurements (\( S_{11}, S_{12}, S_{21}, S_{22} \)) to test whether or not the conversion was accurate. Each waveguide was tested with a 60-90 GHz frequency sweep and the .csv file conversion was done via a MATLAB script. After careful analysis it was concluded to not be within the accuracy of what was needed for phase balancing. Inconsistent phase progression was recorded for each value of the frequency sweep. Within the 37.5 MHz increments, \( S_{11} \) and \( S_{21} \) showed little variation in value; however, each iteration of Equation 3.8 showed at least 10º of phase change. The juxtaposed behavior between the two values was caused the complex \( S_{11} \) value to be inaccurate. Therefore, the NRW method could not be utilized here.
3.2.3.2 Conversion using only $S_{11}$ Short and Open Readings

It was decided that the PDN was too sensitive to apply formulations found in other research. The conversion between $S_{11}$ and $S_{21}$ needed to be specific to the arms of the adapter. After receiving additional information from the PDN’s manufacturer Sage Millimeter Inc., it was understood that the adapter is simply a formulation of bending waveguides. Knowing the length of the waveguide and an open or short $S_{11}$ reading, an $S_{21}$ reading could be accurately predicted. More briefly put, because the adapter is symmetric and reciprocal, an $S_{11}$ measurement from a shorted output shows the phase progression of the signal traveling the length of the channel twice, along with a sign change due to the reflection coefficient $\Gamma = -1$ with $180^\circ$ phase shift (if the DUT is considered to be a PEC). If the phase progression of the signal going forward and traveling back is divided in half, and accounting for the $180^\circ$ shift, it would be theoretically possible to predict the phase at the non-flanged end, i.e. the $S_{21}$ phase value. This relationship can be shown with the following equations [21]

$$\varphi_{21\,short} = \frac{\varphi_{11\,short} - 180^\circ}{2} \pm 360^\circ n \quad (3.9)$$

$$\varphi_{21\,open} = \frac{\varphi_{11\,open} \pm 360^\circ n}{2} \quad (3.10)$$

where

$$n = \text{integer values } 1,2,3,…$$

$$\varphi = \text{S-parameter phase values given by an open or short measurement.}$$

There are two equations because this process is not limited to a shorted measurement; rather, if a quarter wave offset is applied to the short, it may give an open reading [3]. Therefore, it was decided that one or both equations could be used. These two equations are similar to Equations (3.6-3.8), but do not use or require $S_{22}$ measurements. To test these equations, a Keysight ADS
was used to simulate a short and open reading with a 2” waveguide connected to an oscilloscope to read the phase. The simulation showed that indeed the phase of the $S_{21}$ value is half of the $S_{11}$ value.

Figure 3.5. Keysight ADS simulation agreement with Equation 3.9.

Figure 3.6. Keysight ADS simulation results.
To test this method experimentally, a 2” waveguide was connected to a VNA and the S-parameters were measured. The equations gave accurate predictions for both open and short readings as predicted by the simulations. To apply these equations to the adapter, the same procedure was carried out using copper tape as a short. The VNA’s millimeter waveguide extender was connected to the flanged end of the adapter and the copper tape was laid flat on the non-flanged end.

During these measurements the phase reading of the $S_{11}$ began to wander as the adhesive slowly came undone on the adapter. The change was slow but nonetheless was disadvantageous for data repetition and accuracy. It was required that a more secure short be implemented to keep the phase constant. An aluminum disk was fabricated with a center hole to align with the adapter hole so that both may be bolted together; however, the non-flanged end of the adapter was slightly raised around the ring of the output waveguide ports. This caused a not flat connection between the disk and adapter, even when bolted. In addition, using a quarter wave offset required some method of adhesion to place it in between the adapter and the copper tape. A clamp was not a valid option due to the geometry of the adapter. A strong adhesive tape was used in the initial experiments and it worked but was not the most viable option. The use of an aluminum disk also did not allow a quarter wave offset because of the imbalance brought when it was bolted to the adapter.

3.2.3.3 3D Printed Support Structure for $S_{21}$ Readings

With the specific design of the PDN’s adapter (Fig. 2.8), configuring the PDN via $S_{21}$ readings was quickly abandoned due to the second port’s incompatibility to be securely connected to the VNA. However, after the previous two methods proved inaccurate, $S_{21}$ readings were revisited. Therefore, in order to overcome the adapters design and provide repeatable data,
an additional component was required. The method to do so was to incorporate a 3D printed support structure, one that fits tightly to the output end of the PDN’s adapter, Fig. 3.7.

The goal of this new component would be to firmly combine a pin-less waveguide to the PDN’s adapter and secure the connection via bolts and screws that would attach the two pieces together firmly.

The material of the support piece was made with 5\textsuperscript{th} Version Tough Resin that provides a reusable structure that allows for repeated testing. As shown in Figs. 3.8 and 3.9, the support structure, like the PDN adapter, is entirely symmetric. Therefore, four connecting ports were included on the support structure to allow quicker measurements or alternative port connections. The symmetry of the bolts allowed even pressure on the waveguide-adapter connection. This, along with the center bolt, provided a fastened connection that led to repeatable results.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{support_structure.png}
\caption{CAD illustration of the support structure.}
\end{figure}
Figure 3.8. Photograph of the 3D printed support structure.

Figure 3.9a. Photograph of the set-up for $S_{21}$ Measurements with the 3D printed support structure.
Once it proved reliable, the support structure was used to accurately configure the PDN’s adapter, along with the other passive components. Rather than measuring the PDN’s adapter solely with the support structure, the attenuators, phase shifters, and isolators were included to achieve a reading as close to the full setup as possible. The other components, i.e. both sets of power dividers and the waveguides in-between them, had to be measured separately due to the power dividers multiple outputs.

This left four sets of data per channel that added together equaled the full phase progression of the system. It should be noted, however, that this would not be a single-phase reading per channel, but rather a phase range due to the attenuator and phase shifter. Therefore, each channel can be represented as a phase array. The first data set was the channel’s phase progression with both attenuators and phase shifters completely opened (or loosened). Next, the attenuators and phase shifters were completely fastened and the difference between the loosest
and tightest reading was added to the value of the first element, which gave the phase band. The last phase value in the band would be the 2\textsuperscript{nd} bound of the channel phase range, or the last element in the phase range array. With the minimum and maximum of the ranges, the rest of the array can be filled in. The phase band can be represented by:

\[
\phi_{PD1} + \phi_{WG} + \phi_{PD2} + \phi_{Open} = \phi_{ChLwrBnd}
\]

\[
\phi_{ChLwrBnd} + \phi_{PhaseBnd} = \phi_{ChUpperBnd}
\]

\[
[\phi_{ChLwrBnd}, \ldots, \phi_{ChUpperBnd}] = \bar{\Phi}_{FullChannel}
\]

where

\[
\phi_{PD1} = \text{the initial power divider’s phase reading (the PDN’s input)},
\]

\[
\phi_{WG} = \text{phase progression of wave guide that connects power dividers 1 and 2},
\]

\[
\phi_{PD2} = \text{phase progression of the 2\textsuperscript{nd} power divider that feeds into the adjustable components},
\]

\[
\phi_{Open} = \text{phase progression of the rest of the system with phase shifters and attenuators set to open},
\]

\[
\phi_{ChLwrBnd} = \text{the 1\textsuperscript{st} extremity of the channel’s phase range},
\]

\[
\phi_{PhaseBnd} = \text{phase difference between the loosest and tightest reading of the phase shifter and attenuators},
\]

\[
\phi_{ChUpperBnd} = \text{the 2\textsuperscript{nd} extremity of the channel’s phase range}.
\]

With the phase ranges of all 12 channels from input to output, a comparison could be made by investigating if the channels share a common phase value in these ranges. If so, this can be the phase standard to calibrate the PDN. A script was created to handle the calculations and compare the 12 arrays. However, it could be further implemented to perform these calculations for a given frequency value and reiterate the process for each frequency value in the .csv file. Iterating through all frequencies given by the .csv file would constitute large data handling; therefore, it
was determined that Python’s lightweight data manipulation methods would best be suited here. This process was iterated for 80 frequency values between 76 and 80 GHz. If a common value was not found a new frequency would be analyzed. Out of the 80 frequencies tested, 9 were found to share a common value; therefore, there are 9 different options to calibrate the PDN. Each option was carefully considered, and the most viable option was to calibrate the PDN to -150° at 79.35 GHz. In the lab, the PDN was successfully calibrated to where each channel was matched to -150° at the output of the PDN’s adapter. Additionally, calibration was chosen to be -150° because the phase could be matched without a large reliance on the attenuators. After calibration process in the laboratory the largest attenuation difference between ports did not exceed 1 dB. After the PDN was calibrated to feed the Power Combiner with a co-phased and co-attenuation, it was ready to be tested see if the unsymmetric lobes were corrected.
4. Chapter 4: Experiments and Results

4.1 Initial Experiments

As mentioned in Chapter 3, the initial testing without calibrating the PDN caused asymmetric signal input in the Power Combiner, leading to destructive interference and higher order parasitic modal leakage. This imbalance in the input can be seen in the radiation pattern in Fig. 4.1.

![Radiation Pattern](image.png)

*Figure 4.1. Initial tests that showed an uneven radiation pattern.*

In Fig. 4.1, the frequency used was 78 GHz with no prior tuning of the PDN. The reason for the asymmetric lobes was not known at the time; therefore, other solutions were analyzed in an attempt to remedy the field pattern.

In an early experiment, the full system was measured with the different polarizations of the feed waveguide labeled Twister On and Twister Off. Twister On shifted the input waveguide to be vertical with the Twister Off left the feed waveguide unchanged. Rather than a specific frequency, both sets of data were recorded over a 7 GHz band between 75 and 82 GHz with a 35 MHz incremental step. Seen in both data sets, the field behavior shifts to the change in the
operating frequencies within the Power Combiner. The change in frequency was expected to alter the behavior within the system, which then affected the radiated outcome [6]. Twister On and Twister Off are the two biggest data sets collected from the Power Combiner given the initial-untampered PDN. With 201 frequencies tested for each degree between 0° and 360°, the fields changed shape significantly throughout the entire frequency scan. Many patterns did not show any sign of cohesive radiation and were overcome with parasitic configurations, however, some instances showed results that were moderately cohesive. One example of this is shown in Fig. 4.2 for both Twister Off and Twister On. However, these plots were still far from acceptable. Each plot contains five difference frequencies. It should be noted that both plots cover the same frequency range, but do not share any qualities for a given frequency value. In most cases where one showed decent results, the other was poor. In this instance, Twister On and Twister Off were test between 75.350 and 75.490 GHz and displayed different results. For the majority of testing, the patterns proved Twister Off as the better polarization method especially in the lower part of the band. As a note, the idea of constant change in radiation patterns was interesting and lead to further realizations that will be further speculated in Chapter 5.
Figure 4.2a. Radiation pattern for Twister Off experiments.

Figure 4.2b. Radiation pattern for Twister On experiments.

The Twister Off radiation field would have shown a similar result to that of the Power Combiner’s simulated radiation pattern in Fig. 1.4 had it not been for the power abundance in the side lobes. Both main lobes appear to have symmetry along theta and the power levels were close but still unequal (>1 dB).
The data for Twister On in this frequency band showed similar behavior to Fig. 4.1. The modified feed polarization produced a large change such that the radiated fields between Twister Off and Twister On are not comparable for most data points. However, the right-hand side of the main lobe did show similar shapes with a null at 30°. This null is the conjunction between the main and side lobes of the simulated plot. In both figures, the power densities are lower due to the widening of the main lobes, which led to a decrease in power by about -2 dB. This inadequate configuration led to an undesirable main lobe shape; overall, it was incomparable to the simulated results. It should be noted that there were few fields within the data set that were comparable to the simulated results. These fields’ patterns were akin to the simulated fields but were not efficient enough to use for further analysis.

As seen in Fig. 4.2, a general problem in the field plots were the presence of black lobes. Methods used to mitigate the back lobes will be discussed later in the chapter.

It was eventually determined that the initial setup of the PDN was not calibrated to operate at most to any frequencies required i.e. 75-82 GHz. A difference in phase and power could lead to an imbalance in power distribution at the input of the Power Combiner. This could be the reason for many unintended behaviors that would ultimately lead to a deterioration in output. It was deemed necessary to focus a great deal on calibration efforts as described in Chapter 3. Once the system was calibrated to operate at 79.35 GHz, the system was tested again to ensure the main lobes retained symmetry, as seen in the simulated results.
4.2 Field Pattern with Balanced Setup

With the PDN calibrated to provide a balanced phase and attenuation reading at the input of the Power Combiner, the system was ready to be measured experimentally. The full system, PDN attached to the Power Combiner, was placed in an anechoic chamber with a receiving antenna placed at the opposite side of the chamber approximately 3.5 meters away as seen in Fig. 4.3. The Rx antenna was an E-band conical horn antenna with spherical to rectangular adapter. The same microwave instruments used to calibrate the PDN were used for the radiation testing this included the VNA and Rx and transmitting (Tx) millimeter wave extenders. The Tx millimeter wave extender was placed on a rotating place encased in a styrofoam. The full system is shown in Fig. 4.3. A series of E-band waveguides was connected to the millimeter wave extender and the input to the PDN. The weight of the full apparatus was evenly distributed and provided a secure balance on top of the styrofoam. Careful considerations were taken when aligning the Power Combiner and the Rx antenna due to the effective aperture being on the order of millimeters. The conical horn antenna, along with the adapter, were attached to the receiving millimeter waveguide extender.
Since the height of the Power Combiner could not be adjusted, the Rx antenna would have to be adjusted via height calibrator. Finally, the testing environment was ready.

The test was controlled using EM Quest software that measured the radiated field and shifted the rotary table by the degree specified by the user; the initial test was a degree per measurement. After the first test, it was determined that a degree rotation would not account for power that existed between each degree. Therefore, it was determined that a half degree would be the increment used. In Fig. 4.4, it can be seen that the first test contained better symmetrical properties than that of Fig. 4.1. It should be noted that the tick marks between the two are not equal. In Fig. 4.1 the difference in ticks was 2.9 dBm compared to 1.2 dBm in Fig. 4.4. This continued the idea that the PDN’s lack of calibration was the main issue in the output radiation.
Figure 4.4. Radiation pattern of the first test with calibration and horn Rx.

In the following test, the same setup was run once again but with a $0.5^\circ$ rotation as described earlier. The result is shown in Fig. 4.5.

Figure 4.5. Radiation pattern for the second test with $0.5$ degree shift.
The radiation pattern here closely resembles the previous plot and was enlarged to show the minor differences. However, what is different here is the accuracy. The two main lobes reach the same max power level and are very similar in shape. After additional tests were ran, it was concluded that the symmetry in radiated fields was directly proportional to the PDN’s calibration.

4.2.1 Consideration of Back and Parasitic Lobes

One issue shared across all radiation patterns was the presence of noise in the back lobes. During the initial test, an aluminum panel was attached to the back end of the Power Combiner in an attempt to be used as a plane reflector but ultimately showed little improvement [18]. During the new phase of measurements, it was hypothesized that the back lobes were also dependent on the PDN and the waveguides that fed the system. Temporary foam pillars were added to the corners of the PDN ground plate and aluminum foil was wrapped around the system only to expose the Power Combiner. This seemingly quick-fix solution would, however, give further insight into whether the back lobes were a result of radiation leaking from the PDN. Below are the experimental setup and the radiation result. Comparing Fig. 3.6 with Figs. 4.4-4.5, there is a 2 dB improvement which corresponds with an almost 1 dB gain in the main lobes after adjusting the waveguide connections. It can be determined that there is parasitic signal radiation coming from the PDN and the waveguide source.
Figure 4.6. Photograph of the full system wrapped in aluminum foil.

Figure 4.7. Radiation pattern for the aluminum foil-wrapped system.
The amount of signal that leaked can be considerable in critical areas, namely the waveguides that connect the frequency extender to the PDN. To further investigate, one waveguide was poorly fastened to see the effects of how prevalent the leak was.

Figure 4.8 demonstrates a radiation measurement where a parasitic lobe can be seen in between the two main lobes. The entire system remained covered in foil, but the bottom waveguide source was exposed. The new lobe was created simply by unfastening one of the four screws for a signal waveguide connection. This could account for one of the reasons the PDN system, in general, was lossy. Due to the design, there are some components of the PDN connected with only three screws. Although these losses are not substantial, it implies unaccounted perturbations in the system (screws loosening) could cause additional unaccounted loss. However, the loss from the waveguide source feed was substantial and the screws needed to be tightened often. For future testing and experimentation with the full setup, it would be beneficial to find an improved method to connect the millimeter-wave extender to the PDN, rather than using a series of waveguides.

4.4 Analysis of Power Combiner’s Attributes via Experimental Results

With the experimental success of the Power Combiner and PDN together, it was then necessary to compare the Power Combiner’s experimental and simulated properties. The
agreement between the experimental and simulated radiation patterns indicated confidence that
measured and calculated characteristics of the Power Combiner would be accurate. Therefore, to
determine the quality of the physical model of the Power Combiner; the efficiency and gain were
first considered. To find gain, the simplest way was to normalize the radiation pattern of a
laboratory antenna with known gain and compare with the Power Combiner radiation pattern
[23]. Careful consideration was needed before the efficiency could be calculated. The efficiency
of the Power Combiner was determined by comparing the input power to the power received via
a radiation pattern integral, along with the free space loss factor. However, due to the limited
conditions of the experimental setup, the power received was only with respect to the theta
(azimuth) direction. Therefore, the assumption was that the electromagnetic field radiated was
constant with respect to phi, as it is with respect to theta.

4.4.1 Discrete Volumetric Integration Based on a 2D Projection with Symmetric
Properties

The derivational process of this integration method began with the effective aperture area
of the receiving antenna. The effective aperture area of the receiving antenna is characterized as
the area of the antenna that interacts with the incident plane wave, given the far field requirement
[10]. As stated earlier, the Rx antenna was an E-band conical horn antenna; therefore, the
effective aperture area can be considered a circle whose area captures the power density from the
incoming wave from the Power Combiner. This is how the VNA recorded the power received
and the properties of the circularly shaped effective aperture area can be used to obtain insight on
incorporating radiation measurements with respect to the phi direction. This process can be better
understood by Figs. 4.9-4.10. The Fig. 4.9 shows the radiated field with respect to phi, theta, and
Z. The Power Combiner and Rx antenna are not shown, but would be orientated along the Z-
axis. Each blue circle represents a discrete effective aperture reading recorded by the VNA. Each reading spans across the theta plane, i.e. the direction in which the Power Combiner and PDN are rotated. In Fig. 4.11 the distance between aperture readings is visualized as rings with width equal to the radius of an effective aperture area plus the arc length of the 0.5° rotation. These rings are orientated in the phi plane. Therefore, assuming symmetry between phi and theta, one could expand the area of the receiving antenna’s effective aperture area. This expanded area is the ratio of this phi rings to the area of one effective aperture reading. This value, termed “area coefficient,” is multiplied with power received for a given theta reading to find the power received with respect to theta and phi. This process is done for every value of theta and then divided in half due to the symmetry of the lobes [24].
Figure 4.9. Radiation pattern diagram with Rx effective area in spherical coordinates.

Figure 4.10. Radiation pattern in the phi direction.
For clarity, it should be noted the term “rings” is used to indicate they are made of two circles with radiuses $r_{1i}$ and $r_{1o}$, which indicate the radius of the inner circle and the radius of the outer circle that comprise the ring.

As seen in Fig. 4.11, in order to solve this integral the area of the rings that lie in the phi plane (termed phi rings) needed to be calculated. These phi rings are dependent on the distance between theta rotations, as well as the radius of an effective aperture circle. The shifts in effective area were not constant; therefore, the radius and area of the rings would also alter per
measurement (this change in value was dealt with an iterative Python script discussed later in the chapter.) It was assumed that the alignment of the Power Combiner would always be toward the center of the effective aperture circles; this was utilized for two reasons. First, if the rings were defined as the space between leading edges of two adjacent effective aperture circles, the ring radius would be the distance between measurements plus the radius of the effective aperture circle. This meant the distance between measurements could be calculated easily by the properties of a right triangle. To detail this process, the distance between the first two readings, i.e. the formation of the first phi ring will be shown below.

The length \(L_0\) between the Rx and the Power Combiner was known and constant. Next, with theta \((\theta)\) and \(L_0\) known, the distance between the Power Combiner and Rx antenna after the first incremental shift can be found using

\[
L_1 = \frac{L_0}{\cos(0.5)} \quad (4.1)
\]

This distance is also the hypotenuse of the proposed right triangle. There with two of the three sides the last can be found using

\[
r_1 = \sqrt{L_0^2 + L_1^2} \quad (4.2)
\]

where \(r_1\) represents the distance between measurements as seen in Fig. 4.11. To find \(r_1, r_2 \ldots r_n\), using this method, these equations can be modified to become

\[
L_n = \frac{L_0}{\cos(0.5 \ast n)} \quad (4.3)
\]

\[
r_n = \sqrt{L_0^2 + L_n^2} \quad (4.4)
\]

Here:

\(n = \) the number of times the rotating plate has shifted.
Once the distances in measurements were found, the inner and outer radii of the rings can be found using

\[ r_i = r_n + r_{EA} \quad (4.5) \]

\[ r_o = r_{n+1} + r_{EA}, \quad (4.6) \]

where

\[ r_i \] = radius of the inner ring

\[ r_o \] = radius of the outer ring

\[ r_{EA} \] = radius of effective aperture circles.

Therefore, using the radius we can find the areas of the inner and outer circles using

\[ \text{Area}_{i,o} = \pi * r_{i,o}^2 \quad (4.7) \]

The difference between \( \text{Area}_i \) and \( \text{Area}_o \) would give the area of the phi ring. Here, the number of effective aperture circles that could exist within the phi ring, multiplied by the averaged theta power in-between two measurements, given the assumed symmetry between theta and phi, could equal the total power contained in the 3D lobe.

Once the process was understood, it was analyzed with data from a known test between two E-band conical horn antennas in the anechoic chamber. With the characteristics of these antennas known, the radiated and received powers were intuitively understood. These would be the benchmark to test the validity of the integration method, due to a horn antenna’s effective characteristics [22]. A Python script was written to handle the calculations. Python was preferred due to the script’s use of large data processes and iterative methods that differ from MATLAB. For each power reading, the script processed the equations above and once every phi ring was found and multiplied by the received powers with respect to theta the values were summed to give the final value in milliwatts.
Even with the reference test, the confidence of this test was low due the complexity of this calculation, along with the variables that came with it. Therefore, this method of calculating the power received was set aside to utilize other, well documented, methods. A further discussion of this is listed in Chapter 5. Through gain and directivity calculations, the efficiency could still be considered.

4.4.2 Directivity and Gain

4.4.2.1 Directivity

The methods to calculate efficiency could also be used to calculate the gain of the Power Combiner via the directivity. However, due to the novelty of the technique to find efficiency, it was deemed appropriate to use different techniques to find gain. Utilization of common methods could also ensure accuracy if the results agreed with one another. The directivity is only dependent on the directional pattern and can be found via the radiation plots. The radiation pattern measured in Fig. 4.7 suggests that the side lobes are negligible; therefore directivity can be found using

\[ D = \frac{4\pi}{\Theta_1, \Theta_2} \text{ rad, (4.8)} \]

where

\( \Theta = \) Half power beamwidth in 1 of 2 planes (perpendicular to one another.)

However, if the radiation pattern is rotationally symmetric, \( \Theta_{1d} = \Theta_{2d} \) and the equation becomes

\[ D = \frac{4\pi}{\Theta_1, \Theta_2} = \frac{41,253}{\Theta_d^2} \text{ deg, (4.9)} \]

To find the HPBW, the radiation field in Fig. 4.7 was normalized and re-plotted in milliwatts. Seen in Fig. 4.12, the red ring indicates where the normalized value equals half of the max (it is understood that the main lobes in the figure appear to be asymmetric; that is because the plot is
normalized to a spiked-peak maximum from the right-hand side. In addition, the ticks are valued at 0.2, which in the magnitude of received power is small.) The HPBW found from this plot is 17° and using Equation (4.9) gives a directivity value of 18.5 dB.

![Radiation Pattern for 79.35GHz](image)

*Figure 4.12. Radiation pattern in milliwatts with a HPBW ring.*

### 4.4.2.2 Gain

Like directivity, the calculation for gain was possible with a 1D radiation field plot. With respect to gain, the usual power value considered is the maximum value. To use the gain calibration method, the maximum power is recorded between two well-defined E-band horn antennas in the same environment that the Power Combiner was tested. Since the same antenna was used for Rx and Tx, both gain transmitted and gain received were known. Therefore, they could be plugged into the Friis equation as
\[ P_r = \text{Gain}_T + \text{Gain}_R + \text{Loss}_{\text{system}} \implies -30 \text{ dBm} = 2(23 \text{ dBi}) + \text{Loss}_{\text{system}}. \quad (4.10) \]

This gives a loss value of -76 dB. This equation was favorable because it included all the losses in the system, i.e. the free path loss, frequency extender loss, cable loss, etc., all of which applied to the Power Combiner as well. Therefore, with these values, the equation was used again in conjunction with the max power received from power combiner as

\[ P_r = \text{Gain}_{\text{Power Combiner}} + \text{Gain}_R + \text{Loss}_{\text{system}} + \text{Loss}_{\text{PDN}}, \quad (4.11) \]

where \( \text{Loss}_{\text{PDN}} \) is the signal loss from the PDN. These losses are well accounted for from the calibration process, and due to the parallel configuration of the signal division, the total loss of the PDN is considered to be -20 dB. Therefore, the equation becomes

\[ \text{Gain}_{\text{Power Combiner}} = \text{Gain}_R - \text{Loss}_{\text{system}} - \text{Loss}_{\text{PDN}} + P_r \quad (4.12) \]

\[ \Rightarrow \text{Gain}_{\text{Power Combiner}} = -23 \text{ dBi} - (-76 \text{ dB}) + 20 \text{ dB} + (-59 \text{ dBm}) \]

\[ = 14 \text{ dB}. \quad (4.13) \]

From here the equation that relates gain and directivity can be used to find the efficiency.

According to Balanis the relationship is [22]:

\[ G_{re} = e_0 D_0, \quad (4.14) \]

which gives an overall efficiency of \( e_0 = 77\% \). Here the term \( G_{re} \) is the realized gain, where it takes into account mismatch parameters. This value of efficiency the most reliable calculation and portrays a well achieved efficiency given the testing environment.
5. Chapter 5: Discussion and Future Work

5.1 Implications Found from the Calibration

Without a doubt, the largest hurdle to overcome in this project was the calibration process to ensure co-phase and equal amplitude at the input of the Power Combiner. At times, it admittedly felt almost unattainable. As time passed the system was better understood and led to the realization, complexities, and their solutions, and ultimately the discovery of the balanced system. However, what was interesting was not only that calibration was possible (see method in Chapter 3), but additionally PDN configurations existed that could also lead to the same result for difference frequency values. In this high frequency range, changes in the radiation pattern could be seen with small adjustments; when these changes interacted with the PDN’s adjustable components, it was possible to achieve favorable results. This was seen in the results for the initial tests done with the full system without altering the PDN. The Twister On and Twister Off datafile covered the 75-82 GHz frequency band; with this band the radiation patterns for each measurement underwent many changes, some of which led to a radiation pattern similar to the simulated result, as mentioned earlier. This could be seen in the figure below where the pattern showed a strong correlation to what was simulated in Microwave Studios CST, Fig. 1.4.
Figure 5.1. Initial testings’ radiation pattern for a miscellaneous frequency band.

At the time, this was not explored because of results from earlier calibration methods that conceived the idea that a balanced system would be better obtained in the 77-78 GHz range, specifically the frequency value 77.8 GHz. The point of Fig. 5.1 is to show that in the initial state, there could have existed a PDN configuration that would have allowed for a balanced input to the Power Combiner. I believe that either the frequency (in a small variation) or the PDN could be tuned to correct for large side lobes, specifically the phase shifters. The power density at the point of max directivity in Fig. 5.1 is approximately 1 dBm away from the results obtained after calibration. Therefore, this initial setup can be considered to have had little to no attenuation for this frequency.

This led to two underlying ideas, one of which could lead to future work. The first being that this device has a small bandwidth of operation. This is not news as this was realized when
the calibration PDN failed to produce results outside of 79.35 GHz for that particular PDN configuration. The second, however, is that there was a multitude of solutions that do work. These solutions needed to be fined tuned to the proper PDN configuration in accordance with a very specific frequency (specific frequency is relative to the experiment, realistically trends in radiation patterns can be seen up to 10 MHz). This leads to the question of: “Could an optimal configuration exist?” My method of calibration used broad frequency sweep and relied on measured data to find an optimal input for the Power Combiner. For future work, I suggest that a greater emphasis should be made on the radiated fields, possibly by testing the Power Combiner and PDN before calibration, and through radiated field analysis, find a pattern that is close to the simulated results. From here, small adjustments could be made to the system and frequency. Then, analyze the Power Combiners gain, efficiency, etc. to compare with previous results. It should be noted, however, that a proper PDN calibration configuration was found by finding a frequency that works for a given configuration. This is not the case for the opposite procedure, meaning that is no method of finding a configuration that could work for a specific frequency. The best method to combat this would be trial-and-error adjustments after analyzing the radiation pattern for that specific frequency.

5.2 Re-Examining Power Calculations

Due to the nature of the experimental environment assumptions needed to be made to use the radiation pattern for means of the power calculation. These assumptions led to variables that could not be considered which may have led to possible inaccuracies. The variables will be explored here. It should be noted that little to no literature was found for this method of measuring power density with respect to effective aperture. However, the idea remains interesting and I believe that it could be a viable solution with more consideration. Primarily, the
geometry of finding the distances was too vague in comparison to the actual lobe diameter. In other words, the phi rings were dependent on the 0.5° shifts rather than the diameter of the main lobe itself. If the shifts in the rotatory table took into consideration the beamwidth of the main lobe then the results would be more accurate. The area of the effective aperture is very small at this operating frequency; therefore, the 0.5° shifts at that distance would neglect power density that would exist between measurements. In the discrete integral, this was considered and a power average was taken in-between readings to account for these power densities. In retrospect, a small degree shifts would have more beneficial as well. This method was ultimately abandoned due to the time required to verify this was an acceptable method. Furthermore, once the radiated power was found, total losses in the system needed to be accounted for. The consideration of these additional components in the calculation could lead to more errors. This is why the gain comparison was the favorable technique. Utilization of the two reference antennas in the same testing environment as the power combiner would account for the path losses mentioned.

5.3 Future Work

Though a favorable radiation pattern was achieved, there is still additional work that could be done to better the results. As mentioned prior, a method to calibrate the system for a specific frequency would allow better analysis for the desired frequencies. This would also be needed once the frequency was tuned to the higher GHz/THz regime. The current calibration method was proven to work, but there could exist a better alternative. The most crucial test that has been yet conducted is a test that involves a 3D radiation scan. Currently, such a test is being scheduled and planned, but this would be the most reliable option in finding the power combiners true physical characteristics. This would also give the accuracy validation needed for
the discrete integral calculation; this integration process could be used in future testing where the only valid method of measurement would be a 1D scan.
6. Chapter 6: Conclusion

The experimental exploration of the Power Combiner is nothing short of a unique solution to mitigate an otherwise serious and life-threatening situation. The experimental research to validate this design gave reasonable and reliable results that show this is a viable method of signal amplification. The work and results presented in the characterization of the Power Combiner should be continued and verified with better environments to gain further insight of the Power Combiners parameters. There is still much to say about the power distribution network, the novelty of the device sparks interest in all who pass by and the concept serves to further aid in the consideration for the input of the Power Combiner. However, I believe that this system should be used as a steppingstone in conjunction with the Power Combiner. The PDN introduced a considerable number of variables and at times was seen as unpractical. This may serve as a hindrance to the Power Combiner’s overall parameters. With a simpler and more sensitive device the Power Combiner could achieve efficiencies close to what was described in the simulated results, which is required once THz regimes are considered. However, as described before, these tests were a scaled version to the THz range of operation. It is understood that tests ran in the THz range often involve costly devices and practices that deem the scaled version more appropriate and so the use of the PDN was suitable, but these factors still need to be considered. The Power Combiner also serves as an ode to the capabilities of 3D printed technology. This sparks the notion of future devices machined with lab-available components. This alternative has the potential to shift the paradigm of microwave components and the commerce that defines it.
Reference:


