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Passive Front-Ends for Wideband Millimeter Wave Electronic Warfare

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PASSIVE FRONT-ENDS FOR WIDEBAND MILLIMETER WAVE ELECTRONIC WARFARE

by

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has been approved for the Department of Electrical Computer and Energy Engineering

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Dejan S. Filipović

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W. Neill Kefauver

Date_______________

The final copy of this thesis has been examined by the signatories, and we Find that both the content and the form meet acceptable presentation standards Of scholarly work in the above mentioned discipline.
This thesis presents the analysis, design and measurements of novel passive front ends of interest to millimeter wave electronic warfare systems. However, emerging threats in the millimeter waves (18 GHz and above) have led to a push for new systems capable of addressing these threats. Above 18 GHz, traditional techniques of design and fabrication are challenging due to small size, limited bandwidth and losses.

The use of surface micromachining technology for wideband direction finding with multiple element antenna arrays for electronic support is demonstrated. A wideband tapered slot antenna is first designed and measured as an array element for the subsequent arrays. Both 18 – 36 GHz and 75 – 110 GHz amplitude only and amplitude/phase two element direction finding front ends are designed and measured. The design of arrays using Butler matrix and Rotman lens beamformers for greater than two element direction finding over W band and beyond using is also presented.

The design of a dual polarized high power capable front end for electronic attack over an 18 – 45 GHz band is presented. To combine two polarizations into the same radiating aperture, an orthomode transducer (OMT) based upon a new double ridge waveguide cross section is developed. To provide greater flexibility in needed performance characteristics, several different turnstile junction matching sections are tested. A modular horn section is proposed to address flexible and ever changing operational requirements, and is designed for performance criteria such as constant gain, beamwidth, etc. A multi-section branch guide coupler and low loss Rotman lens based upon the proposed cross section are also developed.
Prototyping methods for the herein designed millimeter wave electronic warfare front ends are investigated. Specifically, both printed circuit board (PCB) prototyping of micromachined systems and 3D printing of conventionally machined horns are presented. A 4 – 8 GHz two element array with integrated beamformer fabricated using the stacking of PCB boards is shown, and measured results compare favorably with the micromachined front ends. A plated 3D printed small aperture horn is compared with a conventionally machined horn, and measured results show similar performance with a ten-fold reduction in cost and weight.
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CHAPTER 1

Introduction

The design, development and deployment of effective electronic warfare (EW) systems often have difficult challenges that require exploiting recent advances and innovations. The proliferation of sophisticated electromagnetic defenses and commercially available devices has contributed to the appearance of enemy systems designed to counter current EW capabilities. Whereas narrowband EW systems sufficed in the past, wideband approaches are gaining popularity due to their ability to attack, protect and support various platforms from diverse threats with savings in cost and weight. Traditional techniques for the design of wideband microwave front end systems are generally difficult to use at millimeter wave frequencies due to power handling requirements and losses. Thus, custom transmission lines must be developed for specific application requirements, such as the ability to handle high amounts of continuous wave (CW) power for electronic attack front ends and large single modal bandwidth for electronic support front ends. Modern electronic warfare front ends also must use two orthogonal polarizations, which requires careful design of apertures and beamforming networks.

Conventional approaches to fabricating wideband front end electronic warfare systems at microwave frequencies often fail to produce good and reliable millimeter wave systems. Front-end is a term that generally refers to the circuitry between the radiating aperture and the first intermediate frequency stage in a radio receiver [1]. Herein, it describes the antenna array and beamformer used to feed the antenna elements before active components such as amplifiers or mixers are connected. Hybrid integration of multiple front end components must be minimized to keep losses low. Recent advances in
microfabrication technology allow for the development of wideband front ends operating through hundreds of GHz. High levels of monolithic integration are inherent to most of the proposed technologies. However, many of these fabrication approaches are costly, and thus there is a desire for cost-effective short turnaround prototyping of these systems.

This chapter is organized as follows:

- Section 1.1 outlines an abbreviated background in electronic warfare. Basic overviews of direction finding for electronic support and electronic attack towed decoys are presented as a precursor to topics discussed in the rest of the thesis.
- Section 1.2 discusses common fabrication technologies, including surface micromachining that show great promise for the development of wideband millimeter wave front ends.
- Section 1.3 provides the organization of the rest of the thesis.

1.1 Electronic Warfare

Electronic warfare is considered a critical aspect of modern warfare due to its great effect on a military force’s utilization of systems such as communications, radar, avionics, guided weapons, reconnaissance and intelligence gathering [2] – [10]. As these systems use various portions of the electromagnetic spectrum to perform required tasks, it is critical to ensure that their operations are enabled and protected from any undesirable intrusion or interference. The primary purpose of EW is to gain – as completely as possible – electromagnetic spectrum superiority for electronic signal transmission and sensing. EW can be considered as a synergy of measures that exploit or deny an adversary’s use of the electromagnetic spectrum while defending friendly use of the same. To fight an enemy’s use of the electromagnetic spectrum, EW is used first to obtain information about threats
followed by countering those threats while defending against the enemy’s EW. In general, EW is divided into three branches named electronic attack, electronic protection and electronic support as shown in Figure 1.1.

![Figure 1.1: Subdivisions of electronic warfare and typical activities associated with each branch.](image)

Electronic attack encompasses various applications associated with electromagnetic energy, directed energy, or anti-radiation weapons used to attack specific targets, such as personnel, facilities or equipment, with the intent of degrading, neutralizing or destroying the combat capability of the enemy [11]. Systems built for electronic attack undertake actions to prevent or reduce the efficiency of the enemy’s utilization of the electromagnetic spectrum, employ directed energy weapons, and jam enemy’s radar or electronic command and control. Examples of these systems include:

- Electromagnetic deception using techniques such as false target or duplicate target generation
- Weapons that use directed energy for destruction of enemy electronic systems
- Jamming enemy radar or electronic command and control centers
- Defensive electronic attack systems use the electromagnetic spectrum to protect personnel, facilities and equipment. This includes the use of expendables such as flares or towed decoys, jammers and directed energy infrared countermeasures.

Electronic protection is the branch of electronic warfare that involves actions taken to protect personnel, facilities and equipment from any effects of friendly or enemy use of the electromagnetic spectrum that degrade, neutralize or destroy friendly combat ability [3]. Examples of these systems include:

- Frequency agility in radio communications
- Variable pulse repetition frequency in radar
- Electromagnetic spectrum management
- Electromagnetic hardening of personnel, facilities and equipment

Electronic support is the branch of electronic warfare that involves actions taken to search for, intercept, identify and locate sources of intentional and unintentional radiated electromagnetic energy. This branch often provides a critical source of information for immediate decisions involving electronic attack, electronic protection, targeting and other tactical deployment of forces. Examples of these systems include:

- Radar warning receivers
- Direction finding and geolocation
- Passive and active radar
- Laser warning receivers

A simplified sketch of the electromagnetic spectrum and current capabilities for electronic attack and support is shown in Figure 1.2 [12]. As seen, many current systems operate in either the microwave region or optical region. There is a desire and, even more so, a need to expand the current capabilities of EW systems into the millimeter wave
region. The overreaching theme of this thesis is to address wideband EW front-ends for operation in the millimeter wave region.

Figure 1.2: A portion of the electromagnetic spectrum showing current and desired capabilities for EW [12].

1.1.1 Electronic Support

Radar systems are often used for a variety of electronic support and attack applications. They exploit electromagnetic radiation to be aware of the environment and to determine the range, altitude, direction and/or speed of targets. Passive direction finding systems use the waves radiated from these systems to determine the angular direction of the transmitter. Direction finding techniques come in variety of forms, ranging from simple triangulation based systems to more complex systems that use multiple-element arrays with complicated algorithms for calculating the angle of arrival from multiple beams. The work presented herein focuses on monopulse direction finding systems.

Monopulse direction finding (DF) systems use a comparison between two or more simultaneously received beams to determine the angle of arrival (AOA) of an incoming plane wave [13]. Applications for these systems include radio monitoring, electronic warfare, communications, and remote sensing, just to name a few. The beams of a wideband monopulse DF system can be formed using a single multi-mode antenna [14] or an array of antennas. Typical requirements for a system include high angular resolution, angular sensitivity, range and immunity to jamming. The design of the antenna or array
based front end is critically important to satisfy these requirements. Therefore, the antenna array needs to have reasonable gain/directivity, stable and well calibrated radiation patterns, and low side-lobe levels. Note that the impedance match is not as important as these are receive only systems. However, in this thesis, a concentrated effort is made to keep the VSWR less than 3 over the operating bandwidth.

Typical antennas used for array based direction finding systems include horns [15], dipoles [16], and dielectric rods [17]. Another antenna of interest is the flared notch, or Vivaldi [18]. Vivaldi antennas are endfire low profile planar structures with inherently wide impedance bandwidth. Also, they enable relatively easy integration with beamforming networks. Typically designed as coplanar [19] or antipodal [20] configurations, Vivaldis are traveling wave antennas capable of producing low side lobe levels when the length is a few wavelengths. When operating in an array configuration, mutual coupling between neighboring elements can allow for greater bandwidth, smaller size and improved pattern stability.

Monopulse DF systems typically detect the AOA of an incoming plane wave by creating phase only, amplitude only, or amplitude phase comparisons [21]. Phase only systems, or interferometers, utilize phase differential between two or more antennas. They are accurate, but typically require larger and more complicated back-end systems. Amplitude only systems use two or more beams that are squinted off boresight. They have simple beamforming networks, and the beams can be scanned using phase shifters. The bearing accuracy of an amplitude only system is typically 5 degrees, whereas a phase only system can offer as low as 0.5 degrees of accuracy [22]. Amplitude phase systems combine the features of the two previous systems, providing more accuracy than the amplitude only while being simpler than the phase only systems. They use comparison between sum and
difference beams to determine the AOA. The left-right ambiguity is resolved by utilizing the difference mode phase comparison.

The performance of amplitude phase and amplitude only DF systems depends on both the antenna array and the beamforming network (see Figure 1.3). The simplest DF passive front end consists of a two element antenna array connected to either a 90° or 180° coupler, with typical beam patterns shown in Figure 1.4.

![Figure 1.3: Sketch of a baseline passive front end of a two element monopulse DF system.](image)

Figure 1.3: Sketch of a baseline passive front end of a two element monopulse DF system.

![Figure 1.4: Typical beam patterns produced when a two element antenna array is fed by either (a) a 90° coupler, or (b) a 180° coupler. The radiation patterns are normalized to their maximum values.](image)

Figure 1.4: Typical beam patterns produced when a two element antenna array is fed by either (a) a 90° coupler, or (b) a 180° coupler. The radiation patterns are normalized to their maximum values.
As seen, both beamformers produce two different modes with unique and ideally orthogonal patterns. When the 180° coupler is used, one mode has boresight maximum, whereas the other mode has a boresight null. Squinted beams off boresight are obtained when the array is fed by a 90° coupler. The 180° coupler is suitable for use in an amplitude phase DF system, whereas a 90° coupler is used for amplitude only DF systems.

To determine the AOA of an incoming plane wave, a direction finding function (DFF) is applied. This function is usually based on the unnormalized beam patterns. For the amplitude only system the DFF is derived as the ratio of voltages seen at the output ports of the 90° hybrid, which can be written as

\[
DFF_{Amp}(\alpha) = \frac{|V_{Left}(\alpha)| - |V_{Right}(\alpha)|}{|V_{Left}(\alpha)| + |V_{Right}(\alpha)|}
\]  

(1.1)

Figure 1.5 shows an illustration of this function when applied to the example of Figure 1.4(a). The field of view (FOV) for an unambiguous AOA of an incoming plane wave is 70° ranging from -35° to 35°. This range is determined by the location of the first null (FN) on the opposite side of maximum for each beam. Because this DF system only needs a ratio between the received voltages, it is the simplest.

The amplitude phase system utilizing a 180° coupler is more complex than the amplitude only DF system. The DFF is obtained from the voltage ratio between the difference and sum channels, which can be written as

\[
DFF_{Amp}(\alpha) = \frac{|V_\Delta(\alpha)| - |V_\Sigma(\alpha)|}{|V_\Delta(\alpha)| + |V_\Sigma(\alpha)|}
\]  

(1.2)
Figure 1.5: DFF for an amplitude only system. The unambiguous range is from -35° to 35° off boresight, as indicated.

Figure 1.6 shows an illustration of this function when applied to the example of Figure 1.4(b). A 140° FOV is obtained in the range from -70° to 70° range as determined by the first null of the sum mode. Due to symmetry of patterns, there is an ambiguity of this algorithm if only the amplitude comparison is used. To resolve this, the phase information from the output ports of the 180° coupler is also required. The phase algorithm describing this requirement can be written as

\[
\psi_{\text{Phase/Amp}} = \angle_{\text{Difference}} - \angle_{\text{Sum}} \quad (1.3)
\]

Figure 1.7 shows the phase DFF behavior of the same discussed example. At boresight, a 180° phase change occurs at the boundary between the dual beams of the difference mode. Using this information, the side ambiguity is resolved, and the AOA of an incoming plane wave can be determined accurately using combined phase and amplitude algorithms.
Figure 1.6: DFF for an amplitude phase system. The unambiguous range is from -70° to +70° off boresight, as indicated.

Figure 1.7: Phase DFF of an amplitude phase system, where ±1 denotes the location (left/right) off boresight.

When using only two elements, a basic direction finding system may be obtained. However, these systems can be susceptible to electronic attack, and their relatively small aperture size makes them more sensitive to multipath effects, where the same signal is received from different directions causing a false AOA determination. The use of multiple elements can improve system performance by increasing performance parameters such as
FOV and gain, and also allows for the use of complicated DF algorithms such as maximum likelihood estimation [23], minimum variance distortionless-response algorithm [24], multiple signal classification (MUSIC) [25] and estimation of signal parameters via rotational variance techniques (ESPRIT) [26]. The disadvantage of using multiple element array based direction finding receiver is the more complicated beamformer and back-end needed to accurately compute the AOA. A basic diagram of a four element array is shown in Figure 1.8 and the radiation patterns and computed direction finding functions are shown in Figure 1.9.

![Figure 1.8: Sketch of a direction finding front end of a four element DF front end. Note that more elements can be used at the expense of a more complicated beamformer.](image)

To determine the AOA of an incoming plane wave, a comparison between adjacent ports is needed. Thus, (N-1) DFFs are generated with N beams. To choose which DFF to use, the two beams with the largest amplitudes within the FOV are utilized. This does limit the FOV to ±35°, and can make the system susceptible to lobe jamming if the
designed array and beamformer generate significant lobes. However, greater accuracy can be achieved over the two element systems due to the steeper slope of the DFF and the ability to use more advanced algorithms for determining the AOA. The proposed designs in this thesis use microfabrication technologies to produce wideband monopulse direction finding front ends in both K/Ka bands and W band and beyond.

![Crossover points for beam selection](image)

Figure 1.9: The DFF for a four element monopulse DF system utilizing the amplitude only technique. Thin lines represent the four beams produced from the array, and the bolded lines represent the computed DFF.

1.1.2 Electronic Attack

The need for electronic attack systems with extension into the millimeter wave frequency bands is growing [27]. One particular system receiving increased interest for millimeter wave electronic attack is expendable towed decoys from aircraft and ship platforms. These decoys are employed to provide a spatial separation from their host platform while generating a signal greater in amplitude than the radar return from its platform to seduce a radar guided missile threat and decoy it away. While current capabilities exist up to the Ku band, ongoing threats in the 18 – 45 GHz range has led to
interest in decoys for this frequency range. Key system parameters for towed decoy front-ends include high power capability, broad beamwidth, high gain and sufficient isolation between orthogonal polarizations. In addition, decoy transmitters need to be dual polarized as the platform does not have a priori knowledge of the radar’s polarization. Shown in Figure 1.10 is a typical block diagram of a standalone type towed decoy [10]. The principle of operation is to receive the signal coming from a radar guided missile and transmit proper return waveforms greater than the radar cross section return coming from the host platform. The radar guided missile is thus seduced by the decoy, protecting the host platform.

![Block diagram of a standalone towed decoy](image)

Figure 1.10: Block diagram of a standalone towed decoy. Note that the aircraft only provides the control signal for the decoy, while all other components are located in the narrow body of the decoy itself [10].

Other types of towed decoys can use information gathered by the host platform, which is transmitted through the tow line and directs the decoy to generate and radiate appropriate waveforms to lure enemy missiles. This thesis provides for the development of
a dual polarized wideband front-end to radiate up to 4 kW effective isotropic radiated power at boresight in a towed decoy platform.

1.2 Fabrication Technologies

Printed circuit board technologies used to create passive components and front-ends at microwave frequencies often fail to accurately fabricate micron-sized features needed for millimeter-wave devices. At the same time, losses from commonly used dielectrics increase and may become detrimental to the overall system performance, especially because the dielectric further reduces the structure size. Power handling capability of commonly used planar transmission lines in the millimeter waves is also limited. This thesis considers the methods by which millimeter wave front-ends may be fabricated and easily prototyped and takes this into consideration for the design of representative passive components.

Monolithic fabrication of passive front ends not only decreases the cost by reducing assembly time, but the ability to fabricate passive components such as impedance transformers, filters, and transitions close to the antenna array decreases transmission line and interconnect losses while improving overall system performance. As fabrication technologies mature, the ability to fabricate millimeter-wave components has reliably and affordably improved. This fabrication ability, along with spectrum crowding below millimeter, has led to the development of new millimeter-wave systems. The millimeter-wave spectrum allows data links to be designed with high bandwidths for fast data-rate communications [28]. Low-cost collision avoidance radars are also developing at millimeter wavelengths due to the increased spatial resolution [29]. Other applications using
millimeter-wave systems include non-lethal crowd control weaponry developed near 94 GHz [30], and various targeting and tracking systems [31] – [32].

Amongst several surface micromachining techniques, the PolyStrata™ fabrication process [33] is unique since it can produce rectangular coaxial (recta-coax) lines with multi-decade bandwidth along with traditional waveguide topologies through the sub-millimeter wave range [34]. The process consists of sequential layering of photoresist, copper, and dielectric straps to produce a layered structure as shown in Figure 1.11. The photoresist is then removed through release holes built in the outside walls, leaving an air-filled transmission line. The number of layers and height thereof can vary between different fabrication runs and are preset by the foundry. This leaves an open door for further performance optimization with dedicated wafer runs.

![Figure 1.11: Illustration of the PolyStrata fabrication process with preselected layer heights. *indicates layers designated by the foundry to support straps.](image)

More traditional fabrication processes are also used in the construction of millimeter wave electronic warfare front ends. The use of computer numerically-controlled (CNC) multi-axis machining, a subtractive process, to produce prototypes and millimeter wave components has been generally disregarded due to the increased time and cost over other available processes [35]. To decrease machining time, either the time to traverse
consecutive routing contours must be reduced or the amount of material removed during each pass must be increased. The amount to which the fabrication process will be speed up depends on material due to tool wear [36]. Today's CNC machine designs are well developed with capabilities such as multi-axis control, error compensation for tooling under/over cuts, and multi-process manufacture (e.g. combined mill/turn/laser and grinding machines) [37].

Electrical discharge machining (EDM) is another common material removal process [38]. Its unique feature of using thermal energy to machine electrically conductive parts regardless of hardness has been its distinctive advantage in the manufacture of automotive, aerospace and surgical components. In addition, EDM does not make direct contact between the electrode and the workpiece eliminating mechanical stresses, chatter and vibration problems during machining. The material erosion mechanism primarily makes use of electrical energy and turns it into thermal energy through a series of discrete electrical discharges occurring between the electrode and workpiece immersed in a dielectric fluid [39]. The thermal energy generates a channel of plasma between the cathode and anode at a temperature in the range of 8,000 to 12,000 °C [40] initializing a substantial amount of heating and melting of material at the surface of each pole. When the pulsating direct current supply is turned off, the plasma channel breaks down. This causes a sudden reduction in the temperature, allowing the circulating dielectric fluid to flush the molten material from the pole surfaces in the form of microscopic debris.

A number of EDM variations based on this basic configuration have emerged in the industry to cope with the machining of exotic materials or super hard metal alloys used exclusively in the manufacture of aeronautical and aerospace parts. Wire-cut EDM is one of the more popular variants owing to its ability to machine conductive, exotic and high strength and temperature resistive materials with the scope of generating intricate shapes
and profiles in single pieces not achievable by CNC [41]. It uses a thin continuously travelling wire feeding through the workpiece by a micro-processor, eliminating the need for elaborate preshaped electrodes, which are often required in EDM.

Methods for the timely manufacture of prototypes for design verification and/or modification have recently focused on a number of rapid prototyping techniques (e.g., stereolithography apparatus, solid ground curing, selective laser sintering, laminated object manufacturing, and fused deposition modeling, among others). These techniques are typically additive processes that produce engineering prototypes in minimum possible lead times. One of the more common prototyping processes is to utilize existing capabilities for the fabrication of a printed circuit board (PCB). Up to microwave frequencies, PCBs are often the backbone of electrical devices, allowing the connection of passive, active and embedded devices together [42]. At microwave and millimeter wave frequencies, losses due to the substrate used and cross-talk between adjacent lines limit their use.

3D printing or additive manufacturing is any of various processes for making a three-dimensional object of almost any shape from a 3D model or other electronic data source primarily through additive processes in which successive layers of material are laid down under computer control. Early additive manufacturing equipment and materials were developed in the 1980s. In 1984, Chuck Hull of 3D Systems Corporation invented a process known as stereolithography, in which layers are added by curing photopolymers with UV lasers [43]. He also developed the STL (Stereo Lithography) file format widely accepted by 3D printing software as well as the digital slicing and infill strategies common to many processes today. The term 3D printing originally referred to a process employing standard and custom inkjet print heads. The technology used by most commercially available 3D printers is fused deposition modeling [44], a special application of plastic extrusion. Additive manufacturing processes for metal sintering or melting (such as
selective laser sintering, direct metal laser sintering, and selective laser melting) are also in current use, though they are usually referred to by their own individual names. Currently, the term 3D printing refers only to the polymer technologies in most minds, and the term additive manufacturing is likelier to be used in metalworking contexts than among polymer/inkjet/stereolithography methodologies. This thesis proposes the use of 3D printing for prototyping apertures operating over the electronic attack band of 18 – 45 GHz. In addition, a novel method of prototyping that mimics the PolyStrata micromachining process through stacking PCBs is also discussed for direction finding front ends.

1.3 Thesis Outline

This thesis is organized as follows:

- Chapter 2 discusses the use of surface micromachining to produce wideband direction finding two element antenna arrays for electronic support. A wideband tapered slot antenna is first designed and measured as an array element for the subsequent arrays. Both 18 – 36 GHz and 75 – 110 GHz amplitude only and amplitude phase direction finding front ends are then designed, fabricated and measured.

- Chapter 3 provides the design of multiple element arrays (N > 2) with integrated beamformers for direction finding over W band and beyond. The design and fabrication of these arrays using either a Butler matrix or Rotman lens beamforming techniques are given. Appropriate comparisons and the benefits of each topology are discussed.

- Chapter 4 discusses the design of a dual polarized front end for electronic attack over an 18 – 45 GHz band. The design of a double ridge waveguide bifurcation and turnstile junction for use in an orthomode transducer (OMT) is shown. To provide
greater flexibility in needed performance characteristics, several different matching sections used in the turnstile junction are designed, fabricated and tested. A modular horn section is then developed and directly attached to the small radiating aperture of the OMT. This modular horn is able to be designed for performance criteria such as constant gain, beamwidth, etc.

- Chapter 5 outlines the design of beamformers for electronic attack applications in the 18 – 45 GHz band. A multi-section double ridge waveguide branch guide coupler is developed, and shown to have designed amplitude misbalance of less than 3 dB and phase misbalance of 90° ± 4° over the desired bandwidth. A high power capable Rotman lens is also designed and fabricated. Measured results show a gain greater than 10 dBi and consistent main beam location.

- Chapter 6 details the use of different prototyping methods for millimeter wave front ends. Specifically, PCB prototyping of 3D micromachined systems and 3D printing of conventionally machined horns are presented. A 4 – 18 GHz two element array with commercially available beamformers is first designed to show good performance of the novel prototyping method. A 4 – 8 GHz two element array with integrated beamformer fabricated using the stacking of PCB boards is then shown as prototyping validation of a surface micromachined array. A 3D printed small aperture horn is compared with a horn fabricated via wire EDM, and measured results show similar performance with a ten-fold reduction in cost and weight.

- Chapter 7 summarizes the thesis, its contributions, and outlines some future work directions.
Appendix A details the design of an OMT operating over a 7.5 – 20 GHz bandwidth with attached gain stabilized horn. The front-end system is fabricated using split-block CNC machining, and confirms that the OMT designed in Chapter 4 is scalable.
CHAPTER 2

Two-Element Array DF Front-Ends For Electronic Support

2.1 Introduction

Direction finding (DF) systems often use a comparison between two or more beams to determine the angle of arrival (AOA) of an incoming plane wave [45]. While printed technologies are frequently used for applications up to Ka-band, systems built in the millimeter and sub-millimeter wave bands are typically dominated by waveguides [46]. Recent advances in fabrication technologies allow for direct integration of planar beamforming networks with antennas at these frequencies, reducing their size, piece count, cost and complexity [47] – [50]. In [47], a silicon substrate based system was constructed with a coplanar waveguide (CPW) fed slot-ring antenna loaded by a hemispherical lens. A design in substrate integrated waveguide [48] used a 32 x 32 slot array antenna with 16-way power divider for narrow beams and high gains. A full transceiver design is presented in [49] using Z cut quartz as a microstrip medium. All these designs have narrow bandwidth within a portion of W-band (75 GHz – 110 GHz). Note that the use of substrate, while providing some mechanical stability, may introduce higher loss, sparse component packaging and higher cross-talk due to poor shielding of adjacent circuits.

Air filled rectangular coaxial line (recta-coax) for millimeter wave applications has received increased recent attention because of its ultra-low loss, dispersion, and cross-talk, as well as tight packaging and ability to reduce size by utilizing the third dimension [33], [51] – [53]. In [51], the Electromechanical Fabrication Process, or EFAB™, is used. This
multilayer process involves three separate steps: selectively depositing the first material, blanket depositing a second material and layer planarization. This process is repeated until a nickel air filled recta-coax line is produced. The inner conductor in this process is typically supported by shorting a quarter wavelength section of recta-coax line, limiting the bandwidth of the structure. The SU-8 process [52] uses metal coated SU-8 photoresist to build up layers of a recta-coax line. Similar to the EFAB process, shorted quarter wave stubs are needed to support the inner conductor of the recta-coax line. In [53], the SU-8 process is used to produce a monopole antenna on a silicon substrate. The many developed components and integrated subsystems indicate that recta-coax may be considered as an alternative to traditional transmission line technologies such as microstrip or stripline. Additionally, smaller form factors and wider bandwidths with regards to rectangular waveguide or substrate integrated waveguide can be achieved. Direct integration with MMICs and other active components has also been shown [50] – [54] for fully integrated RF systems.

In this Chapter we demonstrate that wideband two element DF at millimeter waves is achievable with current technology, specifically PolyStrata fabrication. This demonstration is done through the development of both 18 – 36 GHz and 75 – 110 GHz two element DF front ends using amplitude-only and amplitude-phase methods. A wideband tapered slot antenna operating over a 43 – 140 GHz band is first developed as an array element for these front ends. Hybrid couplers are then monolithically integrated to a two element array to complete the front end design.

This chapter is organized as follows:

- Section 2.2 outlines the design of a W band single element tapered slot antenna used as an antenna element for both direction finding subsystems.
• Section 2.3 introduces the two element amplitude-only DF front-end and discusses its theoretical and experimental performance.

• Section 2.4 reviews the design and results of the amplitude-phase DF front-end.

2.2 Millimeter Wave Tapered Slot Antenna

Tapered slot antennas (TSA) at millimeter wave frequencies have been actively pursued since their introduction [55]. In [56], bulk micromachining is used to remove silicon under the diode-fed Fermi (type of geometry) TSA thus achieving effective permittivity needed for antenna operation within a portion of W-band. Due to the process, the antenna operation is limited up to 120 GHz. A coplanar waveguide fed TSA on silicon substrate [57]; microstrip fed TSAs in liquid crystal polymer [58] and low temperature co-fired ceramic [59]; and a TSA with substrate integrated waveguide [60], have also been demonstrated. A rough comparison between these antennas and the TSA shown herein is shown in Table 2.1. As seen, each of these reported designs has less than an octave bandwidth, which is very atypical for a TSA. Note that at these frequencies, even the above mentioned technologies can have high loss, limited performance due to achievable dielectric constants used in typical designs [61], and be fragile due to their small thickness (particularly [56]). Also, monolithic integration with other system components may be difficult. To obtain a TSA with over an octave bandwidth at millimeter wave frequencies, a recta-coax line (RCL) fed exponentially tapered TSA built using surface micromachining is designed and used as a baseline antenna for further subsystem level development.
Table 2.1: The comparison of various millimeter wave TSAs utilizing coplanar waveguide (CPW), microstrip (MS), substrate integrated waveguide (SIW), and recta-coax line (RCL) feeds. Note that gain variation in the RCL model is due to the electrical size of the aperture changing over the wide bandwidth.

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Length [mm]</th>
<th>Width [mm]</th>
<th>Line Type</th>
<th>Gain [dBi]</th>
<th>Bandwidth [GHz]</th>
</tr>
</thead>
<tbody>
<tr>
<td>[57]</td>
<td>7.255</td>
<td>3.243</td>
<td>CPW</td>
<td>8 - 10</td>
<td>75 - 110</td>
</tr>
<tr>
<td>[58]</td>
<td>5.9</td>
<td>4</td>
<td>MS</td>
<td>5 - 10</td>
<td>70 - 92</td>
</tr>
<tr>
<td>[59]</td>
<td>5</td>
<td>6</td>
<td>MS</td>
<td>8</td>
<td>75 - 82</td>
</tr>
<tr>
<td>[60]</td>
<td>30</td>
<td>2.29</td>
<td>SIW</td>
<td>14 - 15</td>
<td>90 - 120</td>
</tr>
<tr>
<td>UCB</td>
<td>8</td>
<td>2.16</td>
<td>RCL</td>
<td>2.8 - 10</td>
<td>43 - 140</td>
</tr>
</tbody>
</table>

The herein designed wideband millimeter wave TSA based on a single element of an 8 x 1 array from [62] is shown in Figure 2.1. A basic configuration from [62] is chosen since it lends itself very well to the PolyStrata process with much greater design freedom to improve performance and mechanical integration for operation in the millimeter waves. A blown-up photograph of a recta-coax to TSA feed transition, release holes, and strata configuration provides additional detail. For the feed, the inner conductor of a 50 Ω RCL is extended across a 0.2 mm slot on layers 7 and 8 of the strata stackup from Figure 1.11. The RCL has a total width of 0.4 mm, inner conductor width of 0.2 mm, total height of 0.3 mm and inner conductor height of 0.1 mm. A rectangular open end emulates high impedance to complete the feed transition. In order to feed the coax straight into element, it is necessary to bend the slot 90°. To keep the total width of the single element small, the slot is meandered to the radiating region, which is different than other reported designs [57] – [60]. For ease of integration and mechanical robustness, the entire 1 mm stack-up is utilized.

To design a well matched TSA with stable far-field over a wide band, a number of design parameters need to be examined. These include corrugations, taper profile, and growth rate. While other geometrical parameters were also studied (such as feed point
location and size of open end), it is seen that their effect is small and thus they are not included here. All simulations are performed using Ansoft HFSS [63]. The effect of corrugations on the performance of an exponential TSA with a growth rate of 2500 mm$^{-1}$ is studied first. It is well known that corrugations have impact on beamwidth, sidelobe levels, and front to back ratio of planar TSA [20]. While initial corrugation values were taken from [64], due to fabrication constraints, their width and periodicity are set at 0.1 mm (0.046λ at 140 GHz). There are a total of 25 corrugations, with a length of 0.4 mm (0.187λ at 140 GHz). As seen in Figure 2.2a, the addition of corrugations increases the gain of the antenna between 2 and 3 dB over most of the band, while having little impact on VSWR. This is expected, as corrugations on tapered slot antennas are shown to affect the currents on the edge of the TSA element [65].

![Image](image_url)

Figure 2.1: Rendering of a designed TSA with a close up photographs of the fabricated antenna features characteristic for both the antenna and the process. Also shown are the corresponding dimensions.

The type of taper used is found to have a significant impact on VSWR and gain of the antenna, as seen in Figure 2.2b. While there are numerous tapers, three specific types are considered: linear, exponential, and Fermi (profile defined by Fermi-Dirac distribution)
Models of each taper profile are shown in Figure 2.3. The width of the slots in exponential and Fermi configurations are proportional to $e^{Rz}$ and $(1 + e^{-Rz})^{-1}$, respectively. Here, $R$ is growth rate and $z$ is axial length. To ensure design guidelines of the foundry are met, the corrugation length of 0.4 mm and growth rate of 2500 mm$^{-1}$ are used for each taper. Width and length of the TSA are kept constant. As seen, Fermi taper has the worst gain. The exponential taper has the best VSWR. Besides superior performance, the exponential taper is also the best mechanically because its edges are thicker over the length of the taper. Thus, the exponential taper is selected for further parametric study and corresponding design (and fabrication).

Figure 2.2: Effects of TSA geometry on VSWR and gain of a TSA. Shown parameters include: (a) corrugations, (b) type of taper, (c) corrugation length, and (d) growth rate.
The length of each corrugation is considered next in an exponential TSA (R = 2500 mm⁻¹). As the length of the taper is increased, the gain at the lower end of the frequency band is improved as seen in Figure 2.2c. However, when the corrugation length is approximately λ/10 or larger, the gain is decreased. VSWR above cutoff frequency of 40 GHz equivalent to a total element length of λ is relatively unaffected by the length of corrugations, with longer corrugations having a relatively smoother slope over the band.

Finally, the growth rate of the exponential TSA is varied, with corrugation length set at 0.4 mm. Mid-band VSWR is unaffected as seen in Figure 2.2d. However, a slower growth rate has better VSWR above 100 GHz while a faster growth rate has better VSWR below 85 GHz. Note that the optimization with included fabrication constraints will yield
better antenna performance; however, this will require a dedicated single user wafer run. Based on the parametric studies shown in this section, a TSA with exponential growth rate of 2500 mm\(^{-1}\) and corrugation length of 0.45 mm is chosen for fabrication for improved VSWR over the entire frequency band and higher gain at lower frequencies.

To characterize the far-field of the designed antenna over the entire frequency range, several RCL to waveguide transitions are needed due to the waveguide test sets for the instrumentation. The rectangular waveguide cross-sections ranging from WR-22 through WR-7 would cover the entire bandwidth. However, fabrication of all these is considered to not be a good use of available wafer's space, especially since in the final application none of these is used as the RCL is directly connected to the back-end circuitry. Thus, a single recta-coax to waveguide transition is designed and integrated with antenna. A WR-10 E-plane probe is designed and monolithically integrated with the coaxial feed as shown in Figure 2.4, limiting measurements to 65 - 115 GHz. While this is beyond the typical W band frequencies, measurements can be obtained because single modal operation is maintained throughout this frequency range. The antenna is epoxied to a PCB holder board for mechanical protection and to ensure the \(\lambda/4\) short-wall distance and proper alignment with the WR-10 feed. An Agilent N5246A PNA with W-band modules is used for S-parameter measurements, while patterns are taken in University of Colorado near-field / far field anechoic chamber. Simulations are performed using both HFSS (FEM) and FEKO (MOM) [66]. Measured and simulated VSWR and gain are shown in Figure 2.5, while measured and simulated E-plane 3 dB beamwidth and E plane beam maxima location are shown in Figure 2.6. Note the VSWR and calibrated gain measurements are only taken from 75 to 110 GHz due to available calibration standards.
Figure 2.4: Photograph of a fabricated exponential TSA with integrated WR-10 E plane probe. The TSA is epoxied to a PCB holder board for protection and proper alignment with a WR-10 waveguide feed.

Figure 2.5: Simulated FEM and MOM overlayed with measured (over W-band) VSWR and gain of the antenna shown in Figure 2.3.

As seen, good agreement with simulations is obtained. Measured W-band VSWR is below 1.7 while the realized gain is 9.5 ± 1 dBi. The discrepancy in VSWR is attributed to the thickness of PCB holder boards (used to create the λ/4 short) being different than specified. The thickness variation shifts the resonance of the E-plane probe. Differences in the gain are attributed primarily to the close proximity of the WR-10 waveguide used for feeding the antenna in the anechoic chamber. However, both the gain and VSWR follow
similar trends to simulated values. Note that simulations show that the gain increases through 90 GHz, following with the slight decrease due to the chosen corrugation length.

Simulated 3 dB beamwidth in the E-plane is above 45° over the entire range, with the W-band measurements indicating this parameter is greater than 40°. It is highly likely the difference comes from the PCB board placed along the antenna’s E-plane and secondary scattering coming from it. H-plane 3 dB beamwidth varies between 110° and 40° over the entire range. Note that the E-plane beam maximum varies around 7.5° from the antenna’s end-fire direction, which is not typical for most TSAs.

![Figure 2.6: Simulated and measured (over W-band) E plane 3 dB beamwidth and beam maxima locations.](image)

To further investigate this behavior, the radiation patterns at 45, 65, 90, 115 and 140 GHz in E- and H-planes are shown in Figure 2.7. As seen, E-plane radiation patterns are squinted across all frequencies. This is determined to be caused by the 90° bend in the slot region. However, the antenna’s physical width is much narrower because of the bend, and is an acceptable design tradeoff. As expected, the H-plane patterns are symmetric across all frequencies. Measured radiation patterns at 65, 90 and 115 GHz correlate well to the simulated data in both planes. Differences are attributed to the presence of the
secondary scatterers next to the antenna which are not modeled due to their large electrical size. There is also beam narrowing in both planes as frequency increases, which is common for TSAs.

Figure 2.7: Simulated and measured (from 65 – 115 GHz) radiation patterns.
2.3 Two Element Amplitude Only DF Front-End Subsystem

Through the use of two tapered slot antennas fed by an appropriate beamformer, a unit cell for a direction finding array can be constructed. By connecting this array to a 90° hybrid, two squinted beams are simultaneously produced. At the output ports of the hybrid, the extracted signal voltage amplitudes are used to compute a direction finding function (DFF), as outlined in Chapter 1. To ensure a well-performing system, amplitude only unit cells operating in both 18 – 36 GHz and 75 – 110 GHz bands are designed, fabricated and measured.

To increase the gain of the 18 – 36 GHz front end, a flat set of corrugations is introduced to the outer edges of each antenna element. After observing promising initial results, the length of each corrugation is individually tuned in a full-wave simulator for the best gain and impedance performance over the desired bandwidth. Note that full wave optimization may be conducted, but was not undertaken at this time. In all, there are 27 corrugations with a tuned length between 0.1 mm and 3 mm. The surface micromachined array is 27.2 mm long and 16.1 mm wide, while the entire subsystem measures 45.7 mm × 24.2 mm. For measurements in the 18 – 36 GHz range, an E plane probe with the dimensions of a standard WRD 180 waveguide is designed and integrated with the coaxial lines. The fabricated device is shown in Figure 2.8. The entire device is mounted on a PCB holder board and fastened to a WRD-180 to coaxial adapter. Note this board is necessary to insure test set compatibility and mechanical integrity of the device during measurements and is not part of the original PCB configuration.
Obtained reflection coefficients and radiation patterns are shown in Figures 2.9 and 2.10, respectively. As seen the reflection coefficient is less than -7.5 dB over the entire bandwidth. Since this is a receive-only application, the relatively higher value of reflection coefficient than what is commonly considered (-10 dB) is tolerated due to the improved bandwidth. Also, note that the reflection coefficients at the two ports are different indicating that there are some differences between the fabricated and designed articles. As expected, main lobes for both beams are squinted and they match well to simulated results. The discrepancies in the results are mainly due to diffraction from the auxiliary PCB holder board which is needed to measure the PolyStrata array, and the WRD-180 E plane probe transition.
Figure 2.9: Measured and simulated reflection coefficients for both ports of the micromachined two element antenna array.

Figure 2.10: Measured and simulated normalized E plane radiation patterns for the micromachined two element antenna array.

The maximum gain is shown in Figure 2.11 and the field of view in Figure 2.12. Simulated gain from 8 to 10 dBi is obtained, while measured gain varies from 6 to 10.5 dBi over the desired bandwidth. Discrepancy is attributed to scattering from the WRD-180 coaxial adapter and some differences between the fabricated and simulated/designed models. Note that the better performance in the real system is expected since the unit cell will not have the adapter so close to the radiating aperture. Also, the continuous
improvement of the PolyStrata process will reduce fabrication errors such as strata thickness, which will further improve the performance of this or similar front ends. While the gain is affected by the presence of the adapter, radiation patterns are relatively unaffected. Thus, measured and simulated field of view track well to one another, varying from 70° to 35° over the band.

Figure 2.11: Measured and simulated gain of the two element unit cell. Note that the measured results include both beams.

Figure 2.12: Measured and simulated E-plane FOV of the two element unit cell.
To design and demonstrate a W-band amplitude-only DF front end, a similar approach as used in the 18 – 36 GHz front end is used. The addition of corrugations to single element tapered slot antennas is seen to have impact on beamwidth, sidelobe levels, and front to back ratio. In a two element antenna array, corrugations are shown to reduce the turn on frequency of each beam mode while increasing gain and reducing the E plane 3 dB beamwidth. Note that corrugations are easily incorporated in the chosen surface micromachining process. The effect of corrugations in the designed antenna array with an ideal 90° hybrid is shown in Figure 2.13.

![Figure 2.13: Comparison between an all metal and corrugated two element tapered slot antenna array fed by 90° hybrid. Corrugations increase gain (and thus the range of the system) while decreasing the FOV for the subsystem.](image)

As seen, the addition of corrugations has little effect on VSWR. The gain is increased as expected (2 – 3 dB over the band), and due to the more directive beams the FOV decreases with frequency. However, the FOV stays above 45°, meaning that only 8 of these devices are needed for a full circular array. Note that the increase in cost when going
from 6 to 8 element circular array is minimal due to the small subsystem size, where many of these components can be fabricated on the same wafer (let alone a batch of wafers).

The coupled line hybrid and two element antenna array with corrugations are monolithically integrated, and connected to a coaxial WR-10 E-plane probe for measurements, as shown in Figure 2.14. For measurements and mechanical protection the subsystem is epoxied to a PCB holder board. The two element subsystem has a total width and length of 8.8 mm and 13 mm, respectively, with the two element antenna array having a width and length of 3.8 mm and 7.7 mm. The slot width is 0.19 mm at each antenna feed and 1.36 mm at the radiating slot, with a growth rate of 2750 mm\(^{-1}\). The 25 corrugations have a width of 0.1 mm, length of 0.47 mm and periodicity of 0.1 mm. Note that corrugations are only put on the outside of the two elements, and not in-between them.

Figure 2.14: Photograph of a fabricated two element DF element with integrated WR-10 E plane probe. The subsystem is epoxied to a PCB holder board for protection and proper alignment with a WR-10 waveguide feed.
An Agilent N5246A PNA with conventional TRL calibration is used for S-parameter measurements. Radiation patterns are measured in University of Colorado’s combined near field/far field anechoic chamber. For this measurement, millimeter-wave extensions were used for both impedance and far-field tests. Measured S parameters are compared with HFSS simulations in Figure 2.15.

![Figure 2.15: Measured and simulated reflection coefficient for both ports of the two element unit cell.](image)

As seen, the reflection coefficient is less than -10 dB over the entire 75 – 110 GHz bandwidth, and measured and simulated responses for each beam track very well. Radiation patterns at 75 GHz, 90 GHz and 110 GHz are shown in Figure 2.16. Cross-polarization levels are at least 25 dB below the main peaks. As seen, the patterns from the beam maximum to the first null on the opposite side of boresight correlate well. It is also noticed that the beams are quite different in comparison to example beams shown in Chapter 1. This is because of scattering effects of the PCB holder board and WR-10 waveguide feeding structure. Because the shape of the radiation pattern determines the
AOA performance, having the FOV for the system similar to trends from simulations without the aforementioned auxiliary structures ensures the design is robust.

![Left Beam and Right Beam Radiation Patterns](image)

Figure 2.1: Measured and simulated normalized E plane radiation patterns for the unit cell at 75, 90 and 110 GHz.

Gain surface maps for E plane radiation patterns are shown in Figure 2.17 to further show the good correlation between simulation and measurement as well as clear and stable beam squint across the W-band.
Figure 2.1: Measured and simulated normalized E plane gain surface maps for both ports of the two-element unit cell over 40 dB range.

The maximum gain from 9 to 11 dBi and E plane 3 dB beamwidth between 30° and 45° are shown in Figure 2.18, while the field of view varies between 45° and 60° and E plane beam maxima location from 12.5° to 25° are shown in Figure 2.19. Measurements and simulations follow similar trends, but vary primarily because of the PCB holder board and WR-10 waveguide feeding structure. These auxiliary structures are not included in simulations because of their large electrical size. It should be noted that the bandwidth of the fabricated subsystem is limited by the WR-10 E-Plane probe. Full-wave simulations
show that the two element array has a return loss above 10 dB from 68 to 132 GHz with gain greater than 8 dBi over the same band. Also, when compared with the K/Ka-band two-element DF array much better agreement between simulation and measurements is seen even though fabrication is more challenging at W-band. The two arrays were fabricated in separate builds, with the W-band being fabricated a year later, thus proving the previous point that the PolyStrata may indeed produce high quality parts. In addition, the waveguide to recta-coax transition is much more robust for the W band design.

Figure 2.18: Measured and simulated gain and E plane 3 dB beamwidth. Note that the measured results include both beams.

Figure 2.19: Measured and simulated E-plane FOV and location of E plane beam maxima. Note that the measured results include both beams.
To further evaluate the usability of the introduced (or similarly built) direction finding unit cell, the effects of manufacturing tolerances are studied through simulations. Previously shown agreements between various measurements and simulations are good indication about the validity of the herein presented analysis. It is known that surface microfabrication techniques have tight tolerances in the x-y plane. However, the thicknesses of each layer can vary up to 10 %, depending on the technique used (for PolyStrata this variation is only a few percent). For the subsystem shown herein, the most critical dimension is the thickness of the inner conductor in the coupling layers. For our design layers S7 and S8 (see Figure 1.11 for the used strata stackup) are combined to constitute the inner conductor layer for the coupling sections. To see the effect of fabrication tolerances on subsystem performance, these layer heights are varied by 10 % around their total nominal thickness of 50 μm. The impact of this variation on the subsystems reflection coefficient, port to port coupling, cross-pol discrimination and field of view are shown in Figures 2.20 and 2.21.

Figure 2.20: Measured and simulated reflection coefficients for both ports.
As seen, the reflection coefficient varies by ± 1 dB over the entire band, while port to port coupling varies by ± 2.5 dB at the beginning and at the end of W-band. However, throughout W-band the match and coupling remain acceptable for the intended application. Gain is relatively insensitive, only changing by ± 0.25 dB. For the two element unit cell, an important parameter that should remain insensitive to process variations is the field of view. This ensures that the computed direction finding function remains unchanged from element to element. As seen, the FOV varies by ± 2.5° over the entire band, which is acceptable for the AOA application if calibrated. Note that for all the studied parameters, the maximum variation occurs for the largest deviation in layer thickness, the monotonic relationship of tolerance to performance variation implies the design is stable. Because micromachining tolerances have improved, the performance between different unit cells is expected to be similar and thus reliably allow for mass fabrication of this front end.
2.4 Two Element Amplitude-Phase DF Front-End Subsystem

Array based DF systems use a comparison between two or more beams to detect the AOA of an incoming plane wave by creating phase only, amplitude only, or amplitude phase comparisons. Phase only systems, or interferometers, utilize phase differential between two or more antennas requiring phase detection, whereas amplitude only systems use power ratio. Amplitude phase systems combine the features of these two systems, providing more accuracy than the amplitude only while being simpler than the phase only systems. They use a comparison between so-called sum and difference beams to determine the AOA. The inherent ambiguity is resolved by utilizing the difference mode phase comparison. By integrating a 180° coupler with a two element array instead of the previously discussed 90° coupler, appropriate beams can be formed for use of the amplitude phase method for millimeter wave direction finding.

Figure 2.22: Photograph of the fabricated two-element amplitude-phase DF front-end.
As before, the first system considered is 18 – 36 GHz amplitude-phase DF front-end, with the difference being the integration of two Schiffman phase shifters with the previously designed 90° hybrid to create the appropriate phase difference between the antenna ports [67]. The surface micromachined array is 32.2 mm long and 16.1 mm wide, while the entire subsystem measures 50.7 mm × 24.2 mm. For measurements in the 18 – 36 GHz range, an E plane probe with the dimensions of a standard WRD 180 waveguide is designed and integrated with the coaxial lines. The fabricated device is shown in Figure 2.22. The entire device is mounted on a PCB holder board and fastened to a WRD-180 to coaxial adapter. Note this board is necessary to insure mechanical integrity of the device during measurements and is not part of the original PCB configuration.

Obtained VSWR and radiation patterns are shown in Figures 2.23 and 2.24, respectively. As seen measured VSWR are less than 5 over the entire band. While this is more than is expected from the 180° design, measured results follow the trends of the simulated system. Radiation patterns follow simulated results better. Note that there is increased rippling in the sum mode pattern, which leads to decreased field of view. The discrepancies in the results are mainly due to diffraction from the auxiliary PCB holder board which is needed to measure the PolyStrata array, the WRD-180 E plane probe transition and increased imbalances in the designed beamformer.
Figure 2.23: Measured and simulated VSWR for both ports.

Figure 2.24: Measured and simulated patterns at different frequencies.
Co and cross polarized gain are shown in Figure 2.25 and field of view in Figure 2.26. Simulated sum and difference gains from 8 to 9 dBi and 3 to 6 dBi, respectively, are obtained. Measured sum mode and difference mode gains vary from 4 to 9.5 dBi and -1 to 6 dBi, respectively, over the desired bandwidth. Discrepancy is attributed to scattering from the WRD-180 coaxial adapter and shifted resonance of the WRD 180 E plane probe. Note that cross-pol discrimination at the beam’s gain maximum is at least 15 dB over the bandwidth.

![Figure 2.25: Measured and simulated realized gains for both ports of the amplitude phase DF subsystem.](image)

As seen in Figure 2.26, the measured field of view for this system does not track well to simulated values. This poor tracking is because of rapid variations in the measured sum mode pattern, leading to ambiguities at the edges of the field of view.
A similar system operating in W band is designed. The 180° hybrid [68] and two element array with corrugations are integrated and connected to a coaxial WR-10 E-plane probe for measurements, as shown in Figure 2.27. For measurements and mechanical protection the subsystem is epoxied to a PCB holder board. The two element subsystem has a total width and length of 8.8 mm and 15 mm, respectively. The slot width is 0.19 mm at each antenna feed and 1.36 mm at the radiating slot, with a growth rate of $2750 \text{ mm}^{-1}$. The 25 corrugations have a width of 0.1 mm, length of 0.47 mm and periodicity of 0.1 mm.
Figure 2.27: Photograph of a W-band amplitude phase DF front-end.

Measured VSWR is compared with HFSS simulations in Figure 2.28. As seen, the VSWR is less than 2.5 over the entire 75 – 110 GHz bandwidth, and measured and simulated responses for each beam track very well. Radiation patterns at 75 GHz, 90 GHz and 110 GHz are shown in Figure 2.29.

Figure 2.28: Measured and simulated VSWR for both ports.
Cross-polarization levels are at least 20 dB below the main peaks. As seen, measured sum and difference mode patterns correlate well to simulated values. There is increased rippling seen in both measured patterns due to scattering effects from the WR-10 waveguide feeding structure. The measured setup is shown in Figure 2.30 to show the closeness of the feed to the radiating aperture. Gain surface maps for E plane radiation patterns are plotted in Figure 2.31 to clearly demonstrate the unique features of the sum and difference beams over the operational bandwidth.
Figure 2.30: Measurement setup for the W-band amplitude phase direction finding array. Note the closeness of the feeding structure to the radiating aperture.

Figure 2.31: Measured and simulated E plane gain surface maps for both sum and difference modes of the W-band amplitude/phase two element front end.
The simulated and measured maximum gain, cross pol discrimination and E plane 3 dB beamwidth are given in Figures 2.32, 2.33 and 2.34, respectively. Measured sum mode gain from 7 to 11 dBi, cross-pol discrimination from 10 to 25 dB, and E-plane 3 dB beamwidth between 25° and 45° are observed. Measured difference mode gain from 5 to 7 dBi, cross-pol discrimination from 20 to 35 dB, and E-plane 3 dB beamwidth between 10° and 40° are also seen. Calculated field of view over W-band in Figure 2.35 varies between 110° and 90°. Because of increased rippling in both sum and difference mode patterns due to the close proximity of the WR-10 feeding structure, measured field of view does not track well to the simulated value. It is expected that due to the good correlation of other parameters that the field of view will track well with simulated values when the system is directly integrated with back end circuitry.

Figure 2.32: Measured and simulated gain for both measured ports.
Figure 2.33: Measured and simulated cross-polarization for both ports.

Figure 2.34: Measured and simulated E-plane HPBW for both ports.

Figure 2.35: Measured and simulated FOV.
2.5 Summary

Two-element DF front ends for wideband direction finding in millimeter waves have been presented in this Chapter. A surface micromachined tapered slot antenna with impedance bandwidth from 43 – 140 GHz is designed, fabricated and measured. A parameter study is undertaken to improve both the VSWR and gain performance of the TSA. In comparison to other TSA antennas in the millimeter waves, the developed TSA has a wider bandwidth while maintaining comparable size and gain. Measured VSWR < 2 and gain > 4 dBi over this bandwidth is observed. E-plane radiation patterns are not symmetrical about the radiating aperture due to the bent slot line feed.

Using this antenna element and the same fabrication process, two element DF arrays using either amplitude only (90° hybrid) and amplitude/phase (180° hybrid) configurations are developed for an 18 – 36 GHz and 75 – 110 GHz bandwidth. Coupled line 90° hybrids are monolithically integrated with a two element TSA array to complete the amplitude only front ends. The 18 – 36 GHz front end has VSWR < 3 and gain greater than 6 dBi for both beam ports. Measured FOV varies from 65° to 40° over the band, and compares well to simulated values. VSWR < 2 and gain greater than 9 dBi is seen for the W band amplitude only DF front end. The FOV for this system varies from 60° to 40° over the band.

Schiffman phase shifters are then added onto the previously designed 90° hybrids to create an amplitude/phase DF front end. The 18 – 36 GHz system measured VSWR < 3 and gain > 4 dBi and -1 dBi are seen for the sum and difference ports, respectively. Ripples in the measured radiation patterns cause the FOV to be less than the simulated FOV, varying from 100° to 45° over the band. Discrepancy is attributed to variations in layers heights.
during the PolyStrata fabrication. The W band front end measured VSWR < 2.25 over the band, and gain is > 7 and 5 dBi for the sum and difference ports, respectively. Rippling is present in the radiation patterns due to scattering from the waveguide feed. Measured FOV is > 30° over the band, and affected by rippling in the radiation patterns.

Measurements of all arrays confirm their suitability for DF over these bandwidths, while having good performance such as excellent VSWR, high gain and low loss. These DF front ends have a wider bandwidth than comparable front ends, and are more easily integrated with back end circuitry compared to waveguide based designs.
CHAPTER 3

Multiple Element Array DF Front-Ends For Electronic Support

3.1 Introduction

As discussed in Chapter 2, interest in millimeter wave applications has increased in recent years. These applications often require the use of multibeam arrays that are capable of wide angle coverage over a broad frequency range [69]. Multibeam arrays are commonly used in direction finding as they can achieve highly accurate estimation of angle of arrival, particularly since they allow the use of DF algorithms such as maximum likelihood estimation [23], minimum variance distortionless response algorithm [24], multiple signal classification (MUSIC) [25] and estimation of signal parameters via rotational variance techniques (ESPRIT) [26].

There are several methods that can be used to generate multiple beams. The most popular topology is the phased array, which uses a network of power dividers and phase shifters to set correct amplitude and phase distributions across array elements and thus generate a beam in the desired direction. However, at millimeter wave frequencies, active phase shifters or time delay units (TDUs) can have high loss [70] – [71] and power dividers are often narrowband [72]. To generate multiple beams over a wide bandwidth, appropriately designed hybrids and phase shifters can be used in a Butler matrix topology [73]. The Butler matrix is an N input × N output passive reciprocal beamforming network that allows for simultaneous generation of multiple beams. Basic properties include isolation between input ports, linearity in phase with respect to the position of the output
and an increment in phase depending on the input used. While Butler matrices have been demonstrated at microwave frequencies, their wideband implementation through W band is extremely challenging, and their beams often squint over frequency for wideband applications. An alternative array feeding method with low loss and wide bandwidth is the Rotman lens [74], which is a planar electromagnetic lens having inherently broadband performance and frequency invariant beams due to its true time delay (TTD) beamforming. Rotman lens have been demonstrated into the millimeter wave frequencies [75] – [76], but a wideband version operating over greater than a full waveguide band has not been reported above 18 GHz.

This Chapter presents wideband surface micromachined Butler matrix and Rotman lens fed arrays, capable of achieving instantaneous bandwidths of more than 50 GHz and 100 GHz, respectively within the millimeter wave ranges. The multi-element arrays were designed and fabricated with the PolyStrata process in mind and are the state of art prototypes.

The chapter is organized as follows:

- Section 3.2 discusses the design and performance of the 4x4 Butler matrix fed end-fire array. Using the millimeter wave array element presented in Chapter 2 and a Butler matrix feed, four squinted beams are produced. The front end has nominal operation from 75 – 110 GHz, with good impedance performance and pencil like radiation patterns.

- Section 3.3 overviews the performance of a monolithically integrated Rotman lens fed five-element TSA array. To generate a wideband lens with low loss over W band and beyond, a double ridge waveguide fed lens is used. The design of novel transitions from double ridge waveguide to both parallel plate and recta-coax is presented. In all,
5 different types of transmission line are monolithically integrated into the same design. The designed Rotman lens has good impedance and radiation patterns from 65 – 140 GHz, and is band limited by the TSA designed in Chapter 2. This frequency range is the widest non-optical Rotman lens currently available in literature.

### 3.2 Butler Matrix Fed Array

Several narrowband Butler matrix fed arrays operating above 60 GHz have been previously reported. In [77], a monopulse comparator network using a Butler matrix fed 2 × 2 patch antenna array was designed for use at 93.5 GHz. While the system provides detection in both azimuth and elevation, the operational bandwidth is estimated to be approximately 5 GHz. A switched beam array for use in commercial WPAN systems at 60 GHz is shown in [78]. The system, fabricated using traditional microstrip technology, uses a 4 × 1 patch antenna array, which limits its bandwidth to 57 – 63 GHz. Finally, a Ka band Butler matrix fed patch antenna array fabricated using SU-8 photolithography is presented in [79]. As with previous designs, the bandwidth is limited for this front-end from 34.5 – 38.3 GHz.

A conventional Butler matrix contains alternating rows of hybrids (90° or 180°) and phase shifters in an N × N arrangement to provide required array phase progressions. To produce 4 squinted beams, a phase progression of 45°, 135°, 225° and 315° between array elements is required. The circuit diagram shown in Figure 3.1 produces the exact required phase progression, and complete fabricated model is shown in Figure 3.2. Note that transmission line cross overs are often considered a bottleneck for planar and integrated fabrications. It is also important to recognize that the output ports should ideally have the
same magnitudes of the split signal. Deviations in ideal amplitudes and phase, referred to as imbalances or misbalances, will detrimentally affect the beams integrity. Specifically, higher side lobes, reduced directivity, increased cross-pol, etc. are often observed. These effects may yield to deteriorated accuracy of the DF front end. The 90° hybrid is a 3 section edge coupled design, giving rise to wide bandwidth and better location of necessary crossovers in the design. As discussed above, two cross-overs are required for this Butler matrix. In PolyStrata, this is realized through recta-coax lines that exist within two separate layers to minimize cross-talk. A single section Schiffman phase shifter [80] is used to generate the 45° phase shift. The final design occupies area of 20 mm × 13.3 mm.

Figure 3.1: Block diagram of a Butler matrix fed array.

For measurement purposes, the recta-coax lines of the Butler matrix are transitioned to WR-10 openings. The Butler matrix fed array utilizes a trapezoidal recta-coax to WR-10 E-plane probe as shown in Figure 3.3. Simulated performance in Figure 3.4 shows a reflection coefficient < -25 dB and insertion loss < 0.15 dB over W band. The antenna element is an exponential tapered slot antenna which individual design and
performance are thoroughly discussed in Chapter 2. Unused waveguide ports are left open to radiate where they achieve a return loss above 15 dB and are terminated with absorber to reduce possible measurement contamination.

Figure 3.2: Photograph of the fabricated (top) and corresponding computational 3D model (bottom) Butler matrix fed array.
Figure 3.3: Computational model of the recta-coax to WR-10 transition. A single 0.5 mm PCB board is used to create the necessary length for the back plane short.

Figure 3.4: Simulated reflection coefficient and insertion loss for the recta-coax to WR-10 waveguide transition with both ports properly terminated.

Note that the rectangular waveguide limits measurements to W band. However, provided a good agreement with simulations is achieved, we argue that the developed array will perform well throughout its (computationally) demonstrated bandwidth. The array is epoxied to a PCB holder board for mechanical protection and to ensure proper alignment with a WR-10 flange, both critical for measurements. All ports have measured VSWR < 2.5 over the entire band, as seen in Figure 3.5. Measured and simulated normalized E plane radiation patterns are shown in Figure 3.6, with seen good correlation. Note that the
beams are not symmetrical across the E plane because of asymmetries in the antenna array. The Butler matrix fed array beams squint as frequency increases.

Figure 3.5: Simulated and measured impedance for all input ports of the Butler matrix fed array.

Figure 3.6: Simulated (solid) and measured (dashed) radiation patterns for all input ports of the Butler matrix fed array throughout the W-band.
To further investigate this behavior, the location of the E plane beam maxima over frequency is shown in Figure 3.7. Outermost beam maxima varies from 65° to 45° and -50° to -35°, while the innermost beam ports stay relatively consistent at ± 15°. A staircase like behavior is noted in beam maxima location measurements of both inner and outer fed beams due to rippling in measured patterns. The measured realized gains are > 7 dBi for 3 out of the 4 ports of the Butler matrix fed array, with the remaining port being greater than 5 dBi as shown in Figure 3.8. Simulated and measured E plane gain surface plots for all beams are shown in Figure 3.9, further confirming desired operation of the Butler matrix fed array.

![Figure 3.7: Simulated and measured E plane maximum beam location for all input ports of the Butler matrix fed array. Note that the spikes indicate presence of a scattering object in the antenna range.](image)
Figure 3.8: Simulated and measured gains for all beams.
Figure 3.9: Simulated and measured gain surface maps over W band for all four beams produced by the fabricated 4x4 Butler matrix. Noticed increased beam directivity with frequency and the presence of sidelobes that may impact the use of this front-end subsystem for the array-based direction finding.

3.2 Rotman Lens Fed Array

Several Rotman lenses operating above 60 GHz have been previously reported. In [81], a 3 beam port × 5 array port Rotman lens built using low temperature co-fired ceramic (LTCC) operating at 60 GHz is shown. While there is no reported data on the loss of the lens itself, the insertion loss in the delay lines only is 1.7 dB. This gives the insertion loss of the lens itself to be much higher. Measurements are shown only over a 57 – 64 GHz band. A 5 × 7 Rotman lens fabricated on a silicon wafer for system on a chip packages is shown in [82]. A standard microstrip design topology is used, with gold plating on high resistivity silicon. Line loss is reported to be 0.6 dB/cm, and the measured insertion loss of
the lens is between 2.5 and 5 dB. A 3 × 5 lens fabricated using thin film dielectric is shown in [83]. The system has insertion loss of at least 5 dB at 60 GHz, and measurements are not shown over a broad frequency range. In [84] a 13 × 9 lens fabricated on Rogers 3003 is shown. Rectangular waveguide is used in [85] to fabricate a 5 × 6 lens. To save space, the WR-10 waveguide is turned 90° from the typical design so that the waveguide is perpendicular to the parallel plate cavity. A gain surface map shown over W band shows that while the inner beam ports have a consistent beam location, the outermost beams change position by up to 10°. A summary of these millimeter wave lenses is shown in Table 3.1. Note that for these lenses, none take advantage of the inherent broadband nature of the Rotman lens.

To achieve a wide bandwidth Rotman lens at millimeter waves, a low-loss transmission line able to work in a single mode is desired. At these frequencies, microstrip lenses can have rather large losses, and thin substrates must often be used, decreasing mechanical stability. Rectangular waveguide based lenses are limited in their operating bandwidth due to the presence of higher order modes, and can be rather large. To mitigate this, the rectangular waveguide can be oriented with the long side perpendicular to the parallel plate region [86]. However, this method does not allow the inherent broadband nature of the Rotman lens to be fully utilized.

To design a millimeter wave Rotman lens over a wide bandwidth, a proper transmission line must be chosen. To avoid high loss at millimeter frequencies, a waveguide based design is chosen. To improve the bandwidth of a rectangular waveguide based lens, a ridged waveguide can be used. In [87], a single ridged waveguide transition to parallel plate is presented, although the design of the full Rotman lens array is not given. To further reduce the size of the waveguide, and operate over more than an octave of bandwidth, a double ridge waveguide Rotman lens is selected.
Table 3.1: Rotman Lens Above 60 GHz Utilizing Low Temperature Co-Fired Ceramic (LTCC), Silicon (SI), Microstrip (MS), Rectangular Waveguide (RWG), and Surface Micromachining (SM).

<table>
<thead>
<tr>
<th>Ref.</th>
<th>Length [mm]</th>
<th>Width [mm]</th>
<th>Line Type</th>
<th>Bandwidth [GHz]</th>
<th>Insertion Loss [dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>[16]</td>
<td>19.6</td>
<td>20</td>
<td>SI</td>
<td>~ 58 – 65</td>
<td>1.5 - 3</td>
</tr>
<tr>
<td>[17]</td>
<td>20</td>
<td>15</td>
<td>MS</td>
<td>~ 55 - 65</td>
<td>&gt; 5</td>
</tr>
<tr>
<td>[18]</td>
<td>31</td>
<td>27</td>
<td>MS</td>
<td>~ 74.5 – 73.5</td>
<td>&gt; 5</td>
</tr>
<tr>
<td>[19]</td>
<td>16</td>
<td>8</td>
<td>RWG</td>
<td>~ 75 - 110</td>
<td>2 – 3.5</td>
</tr>
<tr>
<td>Here</td>
<td>43.8</td>
<td>32.3</td>
<td>SM</td>
<td>65 – 140</td>
<td>0.75 - 5</td>
</tr>
</tbody>
</table>

Because the layers have discrete and foundry preset heights, a symmetric double ridge waveguide operating over 3:1 bandwidth was unable to be designed. Thus, an asymmetric double ridge waveguide is designed. The asymmetric double ridge waveguide, as shown in Figure 3.10, has a height of 0.8 mm, width of 1.65 mm, ridge width of 0.6 mm, top ridge height of 0.35 mm and bottom ridge height of 0.2 mm. As seen by the propagation constant of the first three modes, single modal operation from 55 GHz to 175 GHz is obtained, with attenuation of approximately 0.1 dB/cm over this bandwidth.

Figure 3.10: Dimensions and the propagation constant of the first three modes of the asymmetric double ridge waveguide.
The most critical transition in the designed Rotman lens is the double ridge waveguide to parallel plate. Ideally, the ridges would have an exponential taper into the top and bottom walls along with an expansion of the overall width and height, as in [87]. However, because of the discrete nature of each layer, a pure exponential taper along the top wall cannot be fabricated. Instead, the exponential taper is approximated by the steps in each ridge layer. For the top ridge, the taper has 5 layers, while the bottom ridge has 3 layers. The waveguide width exponentially flares from the initial 1.65 mm to 2.6 mm. The final ridge step extends into the parallel plate cavity 0.5 mm. Release holes are built into the transition to ensure photoresist removal while having no electrical impact on performance due to their small size. The final length and width of the transition are 3.3 mm and 2.6 mm, respectively shown in Figure 3.11, while its simulated performance is presented in Figure 3.12. As seen, the reflection is < -10 dB and insertion loss below 0.75 dB over the 70 GHz to 170 GHz bandwidth. Note that the addition of release holes changes the insertion loss by less than 0.05 dB.

Figure 3.11: Model of the parallel plate to double ridge transition.
The Rotman lens is then designed following conventional approaches, based on geometrical optics [74]. The approach assumes that there is no multi-path interference, and that all energy leaving a beam port is assumed to be coupled to the array contour, leaving a lens with no loss. In practice, these idealizations are impossible to achieve. For the three-foci based approach, there are 3 perfect points on the beam contour where no path length errors, are expected. For all other points, the path length will not be a linear function of the beam position, and deviations occur. The geometry of the array contour \( P(X,Y) \) is defined by the position of the three focal points \( (F_0, F_1, F_2) \) and the width of the array contour \( (2N) \), and is shown in Figure 3.13.
As seen, there are two off axis focal points, F₁ and F₂ that are located a distance F from the origin and an angle α off the axis. The third focal point, F₀, is located a distance G from the origin. The equations for path length equality between a ray passing through the point P(X,Y) and the ray through the origin are given by

\[
(F₁P) + W + Nsina = F + W₀
\]  \hspace{1cm} (3.1)

\[
(F₂P) + W - Nsina = F + W₀
\]  \hspace{1cm} (3.2)

\[
(F₀P) + W = G + W₀
\]  \hspace{1cm} (3.3)

With each of the rays being defined by

Figure 3.13: Geometric model of the Rotman lens.
The next step is to normalize each of the parameters relative to the focal length $F$. These parameters are defined as

$$x = \frac{X}{F}, y = \frac{Y}{F}, n = \frac{N}{F}, g = \frac{G}{F}, w = \frac{W - W_0}{F}$$

Notice that the parameter, $w$, is defined as the transmission line length needed in comparison to the center transmission line when connecting the array contour to the array elements. Using the above defined parameters, and rearranging, equations (3.1) – (3.3) become

$$\frac{(F_1 P^2}{F} = 1 - w - nsina$$

$$\frac{(F_2 P^2}{F} = 1 - w + nsina$$

$$\frac{(F_0 P^2}{F} = g - w$$

While equations (3.4) – (3.6) become

$$\frac{(F_1 P^2}{F^2} = 1 + x^2 + y^2 - 2ycosa - 2xsina$$
$$\frac{(F_2p)^2}{p^2} = 1 + x^2 + y^2 - 2y\cos\alpha + 2xsina$$

(3.11)

$$\frac{(F_0p)^2}{p^2} = (g - y)^2 + x^2$$

(3.12)

To solve for the dimensions of the array contour, equations (3.7) and (3.10) are first used. The two equations are combined to produce the following relation

$$(1 - w - nsina)^2 = 1 + x^2 + y^2 - 2y\cos\alpha - 2xsina$$

(3.13)

$$w^2 + n^2\sin^2\alpha + 2w\sin\alpha - 2\sin\alpha - 2w = x^2 + y^2 - 2y\cos\alpha - 2xsina$$

(3.14)

Because the off axis focal points are located symmetrically about the center axis, equation (3.14) can be separated into two independent equations where $x$ is replaced by $-x$ and $n$ by $-n$. One equation contains only odd powers of $x$ and $n$ while the other contains the even terms. Using the odd terms for $x$ and $n$, we can solve for the $x$ position of the array contour, given as

$$2w\sin\alpha - 2n\sin\alpha = -2xsina$$

(3.15)

$$x = n(1 - w)$$

(3.16)

It can be seen that $x$ is a function of the position of the antenna array, and the transmission line length difference for that position. The even terms for $x$ and $n$ are given as

$$1 + w^2 + n^2\sin^2\alpha - 2w = 1 + x^2 + y^2 - 2y\cos\alpha$$

(3.17)

To solve this, equations (3.9) and (3.12) are combined, resulting in
\[(g - w)^2 = (g - y)^2 + x^2 \quad (3.18)\]

\[g^2 - 2gw + w^2 = g^2 - 2gy + y^2 + x^2 \quad (3.19)\]

By subtracting equations (3.17) and (3.19), and rearranging, the following relation for \(y\) is derived

\[y = \frac{n^2 \sin^2 \alpha + 2w(g - 1)}{2(g - \cos \alpha)} \quad (3.20)\]

Notice that \(y\) is a function of \(n, \alpha, w\) and \(g\). To calculate \(w\), and solve for the array contour, equations (16) and (20) are substituted back into equation (3.17). This results in a quadratic equation that must be solved, and is given by

\[Aw^2 + Bw + C = 0 \quad (3.21)\]

Where

\[A = 1 - \frac{(g - 1)^2}{(g - \cos \alpha)^2} - n^2 \quad (3.22)\]

\[B = 2n^2 - n^2 \sin^2 \alpha \cdot \frac{g - 1}{(g - \cos \alpha)^2} + 2g \cdot \frac{g - 1}{g - \cos \alpha} - 2g \quad (3.23)\]

\[C = g \cdot \frac{n^2 \sin^2 \alpha}{g - \cos \alpha} - \frac{n^4 \sin^4 \alpha}{4(g - \cos \alpha)^2} - n^2 \quad (3.24)\]
By noting that the array contour must pass through the origin, the value of w can be calculated as

\[
    w = \frac{-B + \sqrt{B^2 - 4AC}}{2A}, g \geq \cos \alpha \quad \text{(3.25)}
\]

\[
    w = \frac{-B - \sqrt{B^2 - 4AC}}{2A}, g < \cos \alpha \quad \text{(3.26)}
\]

Thus, for specific values of the design parameters off axis focal angle \(\alpha\), ratio of the on axis to off axis focal length \(g\), the value of \(w\) can be computed as a function of the array position \(n\) through equation (3.21). The values of \(w\) and \(n\) can then be used in equations (3.16) and (3.20) to calculate the array contour of the Rotman lens. With this procedure, there are three perfect focal points, located at \(\pm \alpha^\circ\) and \(0^\circ\).

To minimize phase aberrations along the beam contour that are not at these points, an optimum value of \(g\) must be found. Rotman and Turner [74] used an analog to phase error analysis in Ruze lens and estimated the value of \(g\) to be equal to \(\frac{1}{2} + \alpha^2\). In [88], a geometrical optic error analysis suggests that the optimal \(g\) is equal to \(\cos^{-1}\alpha\). For this design, the value of \(g\) given in [74] is used. There are many different formulations proposed to calculate the beam contour of the Rotman lens [89] – [91]. The circular contour proposed by Rotman and Turner is used for this design.

A 4 input, 5 output port Rotman lens, with a maximum focal angle of \(\pm 30^\circ\), array port spacing of 2 mm, focal length of 12 mm and design frequency of 94 GHz is designed. The proper design of sidewalls are not specified in design equations, but are typically either dummy ports or absorber to prevent interference with the primary beam [92]. For this thesis, dummy ports terminated with absorber are chosen. The final design lens is shown
in Figure 3.14. To ensure the removal of photoresist in the parallel plate region, release holes must be added to both the top and bottom. The release holes are 250 μm squares with a period of 500 μm. Because these square slots are small compared to the wavelength, it is estimated that any radiation losses will be small. To see the effects of the release holes, the lens was simulated over a 60 GHz to 170 GHz band. The percentage of power delivered to the array ports, reflected, coupled to adjacent beam ports, delivered to the dummy ports, and losses due to copper and radiation are shown in Figure 3.15.

As seen, at least 32 % (5 dB insertion loss) of the power is coupled to the array ports over a 65 to 170 GHz frequency band, with 3 dB insertion loss over most of the band for inner and outer beams. In comparison, there is typically 8 – 10 dB of insertion loss in microstrip based designs. As expected, there is insignificant radiation loss when release holes are added to the lens. Less than 4 % of the power is lost through metallic losses in the lens. Most of the power is coupled to the five array ports for both inner and outer fed beam ports. The percentage of power going to the dummy ports is increased when an outside beam port is fed. This can be mitigated through the addition of more array ports (to capture more incident power coming from the beam port) or by changing the design parameters of the lens itself (such as focal length). Reflected power and the amount of power that is coupled to the other beam ports is small compared to the amount of power coupled to the dummy or array ports over the designed bandwidth.
Figure 3.14: Top-view of the Rotman lens model designed here (top), and the field distribution within the lens when the first left port is excited. Notice that all coupling mechanisms are clearly depicted in the figure including the desired to the antenna ports and undesired to other ports present in the lens.

To complete the design of the Rotman lens, the array ports are connected to the antenna array through the use of carefully routed properly phased lines. There are many array elements that may be designed in PolyStrata process (horns, monopoles, slotted waveguide, etc.) that can be monolithically integrated with the designed Rotman lens. To
ensure wideband operation, a tapered slot antenna array is adopted. The basic array element is thoroughly described in Chapter 2. While a rescaling of the antenna is possible to better fit the frequency range of the designed Rotman lens, it was not undertaken at this time. The array has an aperture width of 1.36 mm, with a total width of 7.6 mm and length of 7.7 mm. Note that the array element chosen will have deteriorated performance above 140 GHz.

Figure 3.15: Power budget when release holes are added to the Rotman lens for (top) outside and (bottom) interior fed beam ports.
To connect the Rotman lens to the TSA array, a double ridge to recta-coax transition is needed. Transitions discussed in literature often use a coaxial line that is shorted across the ridge [93] – [95]. These are however rather large in size in order to contain a stepped impedance transformer and mode converter. To decrease the size of the transition, an integrated E plane probe is placed in the double ridge cross section. The probe is held in place by two via straps that enclose the line as it transitions between layers. The double ridge waveguide is terminated 0.375 mm behind the probe. The probe has a width of 0.45 mm, depth of 0.2 mm and height of 0.1 mm. The probe is transitioned to the top three layers, and the top wall is left open to air to better match impedances between the probe and the recta-coax line. By transitioning to the top layers, the recta-coax line can be located within the top ridge before the double ridge waveguide is terminated. The line is then transitioned to the input height of the tapered slot antenna array. The transition has a total length of 1.83 mm and width of 1.85 mm. The lines are then properly phased to the antenna array according to the Rotman lens design equations. The final designed transition is shown in Figure 3.16, with simulated results shown in Figure 3.17.

Figure 3.16: Model of the double ridge waveguide to recta-coax transition.
Figure 3.17: Simulated reflection coefficient and insertion loss of the double ridge to recta-coax transition from Figure 3.16.

As seen, the reflection of the designed transition is less than -7.5 dB over a 65 GHz to 140 GHz band. The insertion loss is better than 1.4 dB, with the majority of the band having less than 1 dB. Note that this transition is designed specifically for the bandwidth of the tapered slot antenna array, and a better configuration over the entire frequency bandwidth of the Rotman lens may be obtained with further optimization.

To characterize the performance of the designed Rotman lens, a custom double ridge waveguide to rectangular waveguide is needed. The rectangular waveguide cross-sections ranging from WR-15 through WR-8 would cover the entire bandwidth of the full Rotman lens. However, fabrication of separate Rotman lens for each waveguide band is considered to not be a good use of available wafer’s space, especially since in the final application none of these is used as the input double ridge waveguide is directly connected to the back-end circuitry. Thus, a single double ridge waveguide to rectangular waveguide transition is designed and integrated with antenna. A WR-10 E-plane bend is designed and
monolithically integrated for each beam and dummy port. When used as a feed and left open to radiate, the transition has a reflection less than -20 dB and -15 dB, respectively. The transition is shown in Figure 3.18 and measured results of back to back transitions are shown in Figure 3.19. The measured back to back transition has a reflection coefficient of less than -10 dB when one port is left open to radiate. Beam ports are transitioned away from the lens region to create enough separation to avoid shorting by the WR-10 feed flange. Impedance and calibrated gain measurements are limited to typical W band frequencies due to available calibration standards. The lens is epoxied to a PCB holder board for mechanical protection and to ensure proper alignment with the WR-10 feed. The full system is shown in Figure 3.20, and measurements and full wave simulations for impedance, far field patterns, gain, E plane 3 dB beamwidth and E plane beam maxima location are shown in Figure 3.21, Figure 3.22, Figure 3.23, Figure 3.24, and Figure 3.25, respectively.

Figure 3.18: Asymmetric double ridge waveguide to WR-10 waveguide transition.
Notice first a good agreement between simulations and measurements obtained over the band. VSWR is less than 3 over a 70 GHz to 140 GHz band, which is sufficient for electronic support needs. The gain for the innermost beams varies from 5 to 13 dBi over the 65 to 140 GHz range, while the gain for outermost beam varies from 5 to 11 dBi. Dips in the gain correspond well to the frequencies where there is an increase in power delivered to the dummy ports. The 3 dB beamwidth varies from ~40° at 60 GHz to ~20° at 140 GHz. As seen, there is a narrowing of the 3 dB beamwidth over frequency due to the inherent nature of the tapered slot antenna array. Measured values are within ± 5° of the simulated value. E plane beam maxima location is consistent over frequency and measured locations are within ± 5° of simulated values. Measured beams located at 30° and 10° do not conform exactly to the simulated response due to misalignment during measurement and the close proximity of the WR-10 flange to the radiating aperture.
Figure 3.20: Fabricated and simulated model of the complete Rotman lens. Also shown is the US Quarter to visually demonstrate the physical size of the Rotman lens and other integral components discussed throughout this section.

Figure 3.21: Measured (over W band) and simulated VSWR of the fabricated and designed Rotman lens.
Figure 3.22: Measured and simulated radiation patterns for the Rotman lens.

Figure 3.23: Measured (over W band) and simulated gain.
To further show the stability of the beam over frequency, measured and simulated normalized gain surface maps for the beam pointed at 30° and 10° are shown in Figure 3.26. As seen, the plots are similar and confirm the frequency invariant beam location.

Figure 3.24: Measured (over W band) and simulated E plane 3 dB beamwidth.

Figure 3.25: Measured (over W band) and simulated E plane maxima beam location.
3.4 Summary

A millimeter wave 4x4 Butler matrix fed DF array and Rotman lens with 4 input ports and 5 output ports utilizing double ridged waveguide ports have been designed and fabricated. DF required beams are demonstrated computationally and experimentally over extremely wide instantaneous bandwidths. The Butler matrix fed array has measured
VSWR < 2 for all beam ports, and gain greater than 6 dBi over W band. Proper radiation patterns are generated for DF, but squint inwards with frequency due to increasing directivity of the integrated TSA array. In comparison, the Rotman lens fed TSA array has frequency invariant beams, with measured VSWR < 3 and gain > 6 dBi over a 65 – 140 GHz bandwidth. The Rotman lens design incorporates a double ridge topology to reduce insertion loss to less than 5 dB, and monolithically integrates 5 transmission lines for measurement. To the author's best knowledge, this design demonstrated the widest frequency range of a non-optical Rotman lens. Results of this research clearly indicate that the surface micromachining can indeed enable the realization of wideband and high quality array-based DF front ends throughout the millimeter wave (and possibly even in THz) frequencies.
CHAPTER 4

Dual Polarized Front End for Electronic Attack

4.1 Introduction

As discussed in Chapter 1, electronic warfare can be broken down into electronic support (ES), electronic protection (EP) and electronic attack (EA). Electronic attack is the use of electromagnetic or directed energy to attack personnel, facilities, or equipment [10]. There are five common sub-divisions of EA: jamming, deception, directed energy, anti-radiation missile and expendables. Of particular interest for this thesis research are expendable countermeasures, which are deployed from a host platform and perform self-protection functions. The three most common expendable countermeasures types are chaff, flares, and towed decoys. Chaff is dispensed in bundles of thin conductive elements designed to reflect RF energy and disrupt enemy radar. Flares are designed to protect aircraft from IR-directed threat systems by providing a more attractive target to the missile seeker than the targeted aircraft. Towed decoys attempt to provide a more attractive target than the platform they protect.

Modern monopulse radars employed on radar guided missiles use advanced radar techniques and efficient electronic protection to reduce the effect of EA jamming and deception [96]. Towed decoys are used as the last layer of defense against RF-guided weapons. They may depend on their host platform or be entirely autonomous so onboard electronics must be capable of acquisition, processing, response generation and radiation [97]. Common topologies of towed decoys include the repeater, standalone and transmit
decoys. These are shown in Figures 4.1 – 4.3, respectively. The simple repeater model aims to retransmit the targeting radar waveform at a higher signal level than the return from its host platform. The system typically only requires power and control from the host platform. When this configuration receives an appropriate RF signal from the target, it amplifies and retransmits the same signal in hopes of seducing the threat.

Figure 4.1: Block diagram of a towed decoy repeater topology [98].

Figure 4.2: Block diagram of a towed decoy standalone topology [98].
The standalone model is similar to the repeater topology – however, there are two transmit antennas onboard to transmit the received signal in all directions. In addition, onboard signal processors allow the standalone decoy to transmit waveforms other than the one received in order to deceive the target. A transmit decoy relies upon receivers and signal generators on the host platform. The signal is then passed via the tow line to the decoy, which amplifies and transmits in hopes of luring the target. The power supply can either be located on the decoy or the host platform.

![Block diagram of a towed decoy transmit topology](image)

Figure 4.3: Block diagram of a towed decoy transmit topology [98].

The performance of a repeater or standalone decoy can be defined in two ways. The first is the jam to signal ratio, which is given as

$$
\frac{J}{S} = \frac{G_{DR} G_{DT} G_{DA} \lambda^2}{4\pi \sigma L_p^2}
$$

where $G_{DR}$ is the decoy receiver antenna gain, $G_{DT}$ is the decoy transmitter antenna gain, $G_{DA}$ is the decoy amplifier gain, $L_p$ is the polarization loss and $\sigma$ is the host platform’s radar cross-section. Note that the ratio is range independent [98]. However, this assumes
that the decoy amplifier can supply more and more amplification as the target radar power gets larger and larger. In reality, the decoy amplifier will reach saturation at a certain point. As the target radar gets closer, the same maximum power level will be output by the decoy. When operating in saturation, equation 4.1 becomes range dependent and is given by

\[
\frac{J}{S} = \frac{4\pi R^2 P_{\text{Max}}}{P_T G_T \sigma L_p}
\]  

(4.2)

Where \( R \) is the range to the missile, \( P_{\text{max}} \) is the maximum decoy power output, \( P_T \) is the target radar’s transmitted power, and \( G_T \) is the target radar’s antenna gain. Therefore, there will be a certain range from the target radar at which the decoy starts to operate in saturation. In addition, there is a certain range at which the decoy power will be less than the host platform’s target return echo, and thus will not be effective. This is known as the burn through range [98].

The decoy’s low profile and increasing need to cover wider portions of the spectrum pose significant challenges for decoy design and engineering. Not only must the transmission line be able to have a wide single modal bandwidth to handle various target radars, it must also handle high amounts of power needed to properly seduce the target. The towed decoy front end ideally is dual polarized, as the target’s polarization is unknown. The dual polarized front end must often fit in a confined space, and radiate appropriate waveforms.

In this Chapter we demonstrate a wideband dual polarized front end for towed decoy electronic attack. To support the required bandwidth and high power handling necessary, a custom double ridge waveguide cross section is developed. An orthomode transducer (OMT)
with small aperture horn is then developed, along with several variations of the necessary turnstile junction. To provide additional functionality, the use of a modular horn section is then investigated for use with the designed OMT.

This chapter is organized as follows:

- Section 4.2 outlines the design of an 18 – 45 GHz OMT for use in towed decoy applications. The design of a custom double ridge cross section, along with an $E$ plane bifurcation and turnstile junction required for the OMT is presented. To provide additional capabilities, several different matching cones in the turnstile junction are also fabricated. The OMT is fabricated using CNC split block machining, and each matching cone section is measured and compared to simulations.

- Section 4.3 presents a modular horn extension for the designed OMT. A parameter study is undertaken to determine topologies for constant boresight gain and constant $E$ and $H$ plane 3 dB beamwidth. Three separate constant gain horns are chosen from the contour plots for fabrication and measurement.

### 4.2 Orthomode Transducer

Currently available towed decoys operate up to 18 GHz [99] – [100]. To counter emerging threats, an extension of the towed decoy into the 18 – 45 GHz range is desired [27]. In order to cover a wide bandwidth in the millimeter waves that is capable of handling high power, the natural transmission line choice is waveguide. However, traditional rectangular waveguide has insufficient single modal bandwidth, so ridges must be introduced into the waveguide cross section. Ridges help increase the bandwidth by
reducing the cutoff frequency of the operation mode (typically TE\textsubscript{10}) while maintaining a relatively unchanged cutoff frequency of the first higher order mode.

There is a commercially available double ridge waveguide, WRD-180C24, which operates over most of the desired bandwidth [101] with second and third order modes propagating around 45 GHz. Therefore, a custom design with an increased single mode bandwidth at the expense of higher loss is developed [102]. A brief overview of the waveguide cross-section and its dimensions are given in Figure 4.4.

![Waveguide Cross-Section Diagram](image)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>WRD180C24</th>
<th>WRD1845</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bandwidth</td>
<td>18-40 GHz</td>
<td>18-45 GHz</td>
</tr>
<tr>
<td>Geom. Para. ([w, h/w, g/w, w_{ridge}/w])</td>
<td>[7.3 mm, 0.47, 0.20, 0.25]</td>
<td>[6.7 mm, .45, 0.155, 0.24]</td>
</tr>
</tbody>
</table>

(a)

![Mode Propagation Diagram](image)

Figure 4.4: (a) Cross section and parameters of the custom double ridge waveguide, (b) comparison to the WRD-180C24 cross section and (c) propagation of the first 3 modes for each WRD. Note that the propagation constant of the first mode is the same for both cross-sections, while the second order mode for the custom WRD is moved to 50 GHz.
As seen, to increase the single mode bandwidth, the ridge gap is decreased. To provide the same group velocity dispersion as WRD-180C24, the ridge width, waveguide height and waveguide width are adjusted through parametric study. With the final cross section, single mode bandwidth from 18 – 50 GHz is supported, and over 2 kW continuous wave power (with a safety factor of 4 included) can be handled [102]. Note that fillets at the ridge and waveguide corners are chosen to be the same as WRD-180. Larger fillets will increase power handling abilities, but can start to affect electrical performance if they are too large [102].

Orthomode transducers (OMTs) are key components in dual-polarized antenna feeds for applications such as communications [103], remote sensing [104] and radio astronomy [105]. OMTs are four-electrical port devices used to separate/combine two orthogonal polarizations in a common port. OMTs or similar structures have been in use since the 1950s [106]. The major types of waveguide OMT are the turnstile junction [107], folded turnstile junction [108] – [110], double-folded turnstile junction [111], and the coupler-based design [112] – [113]. Of the OMT designs found in available literature [114] – [122], there have been no high-power capable double-ridge waveguide designs found covering the 18 – 45 GHz band, as shown in Figure 4.5. Such a wideband device is a key enabling technology for electronic attack applications in the desired band (or other scaled bands). This section investigates implementation in waveguide to maximize power beyond coax while not losing the associated bandwidth.
The turnstile junction based OMT [107] is chosen as the basis for developing the double ridge cross-section based design. A model of the full system is shown in Figure 4.6. Note that the system described herein is designed for dual linear polarizations. To obtain circular polarizations, a 90° hybrid must be connected to the orthogonal inputs with phase matched lines. A design of a hybrid capable of this is detailed in Chapter 5. As seen, the OMT consists of four basic components/parts: a double ridge waveguide bifurcation, single ridge turnstile junction, small aperture horn and necessary E plane bends in both single and double ridge waveguide lines. The design of the bifurcation and turnstile junction are examined in detail.

The double ridge bifurcation is inspired by typical waveguide based designs. For rectangular waveguide designs, the input rectangular waveguide is expanded to provide a better impedance match, as the two output ports are full height rectangular waveguide. To preserve the compact size needed for the towed decoy application, a different approach is undertaken.
Here, the double ridge waveguide is split in the E plane into two reduced height single ridge waveguide cross-sections. Each single ridge cross-section has the same impedance as the double ridge cross-section, which provides a good impedance match and a low power loss. A model of the double ridge bifurcation and the obtained match and loss are shown in Figures 4.7 and 4.8, respectively. The final dimensions of the bifurcation are a length of 4.08 mm, width of 8.15 mm and height of 6.73 mm. As seen, the reflection coefficient is better than -30 dB over the band, with insertion loss of less than 0.05 dB. Note that the output ports are curved away from each other so that the design of an additional single ridge waveguide E plane bend is avoided.
The single ridge turnstile junction combines the two polarizations into a common square ridge waveguide. The square quad ridge waveguide output is carefully designed to have the same impedance as the single ridge waveguide, and has a width of 5.56 mm and ridge gap of 2.02 mm. To provide a better match, a stub is located at the base of the quad ridge waveguide.
waveguide. Without a matching stub at the base of the turnstile junction, the impedance match of the two orthogonal ports is poor as seen in Figure 4.9.

![Figure 4.9: Simulated magnitude of S parameters for a turnstile junction with no matching cone.](image)

The design of this stub is typically a circular cone for waveguide based designs [107]. However, for this design, several different matching stubs are considered. They are all shown in Figure 4.10. In order, the matching stubs considered are named corrugated cone, cone, double pyramid, swept pyramid and pyramid. Each matching stub has the same ridge blend radius from single ridge waveguide to quad ridge waveguide of 1.5 mm. The corrugated cone has a base of 3 mm and height of 2.5 mm. There are 0.25 mm sections cutout at 0.375 mm and 1 mm from the center. The cone matching stub has a base of 3 mm and height of 2.5 mm. The double pyramid stub, as the name implies, has a second pyramid placed upon the top of another pyramidal section. The bottom pyramidal section has a square base of 3.5 mm and height of 1 mm. The top pyramid has a square base of 1.5 mm and height of 1.5 mm, making for a total height of 2.5 mm for the structure. The swept pyramid uses a constant ridge gap separation throughout the entire stub section. Thus, a
0.521 mm separation in the matching stub is maintained with the single ridge to quad ridge transition. At the top of the transition the width is the same as the ridge width, 1.626 mm. The bottom of the stub transition has a width of 2 mm. Unconnected sections are then adjoined together to create a solid stub. Finally, the pyramid stub has a base of 3 mm and height of 2 mm. Note that each matching stub comes to a point except for the swept pyramid model, which comes to a flat top.
Having designed the turnstile junction and bifurcation, the only remaining task in assembling the full OMT is to integrate the bifurcation structure with the orthogonal port. A positioning conflict arises if the bifurcation structure is simply copied and rotated by 90°, so lateral repositioning of both bifurcations is necessary, which introduces feed asymmetry.
Figure 4.11: S parameter performance for different matching stub sections used in the single ridge waveguide turnstile junction. Shown are the (a) corrugated cone, (b) cone, (c) double pyramid, (d) swept pyramid and (e) pyramid matching stubs.
This asymmetry can be compensated assuming that the E-plane bends occurring between the bifurcation and turnstile have very good match. Since these bends have reflection coefficient below -30 dB across the entire band, their effect was found to be negligible. The compensation then consists solely of routing additional path lengths in the turnstile feeding waveguides to equalize their length. The simulated S parameters are shown for each matching stub in Figure 4.11. As seen, each matching stub has an impedance match better than -10 dB over the bandwidth, and isolation better than 50 dB for the orthogonal ports. Note the differences in impedance match for the two input ports is due to different lengths of transmission line and bends coming into the turnstile junction from the bifurcation.

To complete the design, the double ridge feed lines are transitioned out to create enough space to attach dual WRD-180 coaxial adapters affixed to WRD-180 to WRD-1845 transitions [123]. A small radiating aperture is then added to the square quad ridge horn output [124]. The final size of the device is 60.96 mm × 60.96 mm × 71.58 mm, the majority of which is fixturing overhead. The OMT is constructed using quick turnaround split-block CNC machining, and held together using screws. To allow for testing of several different matching stubs, the complete OMT is split up into 3 construction blocks: OMT, matching stub and radiating aperture. The OMT section is 41.64 mm long, and contains the fixturing overhead to the double ridge waveguide to coaxial adapters, bifurcation and transition to the turnstile junction. The radiating aperture section is 29.94 mm long, and contains half of the turnstile junction and the aptly named aperture horn. The matching stub section contains the stub and bottom section of the turnstile junction, along with single ridge section which allows for the matching stub to be securely fastened to the OMT. A computer model of the design is shown in Figure 4.12. Each matching stub section is shown in Figure 4.13, and the fabrication process in Figure 4.14.
Figure 4.12: Computer model of the complete OMT. Note that the matching stub section is able to be interchanged easily, depending on performance characteristics required.
Figure 4.13: Fabricated matching stub sections used in the double ridge OMT. Shown are the (a) corrugated cone, (b) cone, (c) double pyramid, (d) swept pyramid and (e) pyramid matching stubs.
Figure 4.14: Fabrication process for the split block OMT. (a) The OMT part of the device is fabricated in 8 separate pieces. (b) The ridge sections are affixed to the waveguide sections using screws and aligned with pins, and two separate sections are made. (c) On one OMT section, the matching cone section is aligned, and then the entire OMT section is screwed together on both sides. (d) A similar process is undertaken for the radiating aperture section, which is made of 4 pieces. (e) The radiating aperture and OMT/matching stub sections are then aligned and affixed together using screws.

An Agilent N8363B VNA is used for S-parameter measurements, and obtained reflection coefficient and isolation are compared to simulated values in Figure 4.15 for each matching stub. Note that this VNA is limited to a maximum frequency of 40 GHz. As seen, the measured reflection coefficient is less than -9 dB over the 18 – 40 GHz range for all matching stubs. Measured isolation is better than 35 dB over the frequency band. The discrepancy in isolation is probably due to the small air gaps between the aluminum blocks and non-symmetrical features in adjoining split block pieces. The effect of these gaps can be minimized through the use of a gasket material or mechanical lapping of the mating surfaces, while the non-symmetrical features can be overcome through the use of finer CNC machining or alternative fabrication method. While doable, these would increase the initial prototype cost and was not actively pursued at this time.
Figure 4.15: Measured magnitudes of S parameters for the full OMT front end system with different matching stub sections. Shown are the measured performance with (a) corrugated cone, (b) cone, (c) double pyramid, (d) swept pyramid and (e) pyramid matching stubs.
Each OMT is measured in University of Colorado’s near field/far field anechoic chamber. Measured and simulated E and H plane radiation patterns for each matching stub OMT are shown in Figure 4.16. As seen, there is some discrepancy in the patterns between simulation and measurement. For the wide beamwidth of the small aperture, this is likely due to scattering effects from the mount. The discrepancy lessens as frequency increases and the directivity of the horn is larger.

Measured and simulated gain and E and H plane 3 dB beamwidths for the corrugated cone stub are shown in Figure 4.17 and 4.18, respectively. As seen, measured gain is between 6 and 11 dBi, with both ports differing by less than 0.5 dB over the band. The difference between simulated and measured gain is less than 1.25 dB. Measured beamwidths do not compare well to simulated values in the lower end of the frequency band as seen in the radiation patterns in Figure 4.16(a).
Figure 4.16: Measured E and H plane patterns of the full OMT front end system with different matching stub sections. Shown are the measured patterns with (a) corrugated cone, (b) cone, (c) double pyramid, (d) swept pyramid and (e) pyramid matching stubs.

Figure 4.17: Measured gain for both polarizations of the full OMT front end system with corrugated cone matching stub section.
4.3 Modular Horn Design

As seen in the measured results above, the compact radiating aperture has a wide beamwidth and lower gain than typical ridged waveguide horn antennas. While this particular aperture is sized for fitting in a 1.8” diameter airborne towed decoy platform, other platforms may not have such size restrictions. These platforms may require different radiating properties of the dual polarized OMT. To address this need, a modular horn extension is proposed to fit onto the end of the small radiating aperture presented previously. This modular horn extension may be designed for radiation parameters such as constant gain, constant E and/or H plane beamwidth, etc. Previous studies on quad ridge horns for parameters such as constant beamwidth used the entire length of the aperture for their parameter studies [125]. In this design, the modular horn extension is directly attached to the small radiating aperture, which allows for reconfiguration of the towed decoy depending upon mission parameters. This prevents a smooth transition between the two sections, but desired radiation characteristics are achievable through careful design.
There will be increased rippling in the E plane due to this abrupt transition as compared to a smooth case.

Figure 4.19: Demonstration of the modular horn section designed to be directly affixed to the radiating small aperture of the double ridge OMT. A parameter study that varies the flare angle, flare length, and ridge exponential taper rate is undertaken to see the effects on the far field performance.

A model of the proposed horn extension is shown in Figure 4.19. As seen, there are three key parameters that affect the profile – and thus the far field characteristics – of the quad ridged horn. These are the flare length, flare angle and exponential ridge taper rate. For the study, flare angle is varied between 20° and 60°, flare length between 0.8” and 2” and exponential ridge taper between 1 mm⁻¹ and 101 mm⁻¹. The far field parameters studied are the variation of the boresight gain and E and H plane 3 dB beamwidths. For each of these parameters, the value is extracted every GHz from 18 to 45 GHz. The mean and standard deviation is then calculated for each specific value of flare angle, flare length and exponential ridge taper. Note that the waveguide walls are linearly flared to the
radiating aperture. They may be adjusted to an exponential taper in the future. This is plotted in contour plots to aid in the design choice. Contour plots for constant boresight gain are shown in Figure 4.20.

Figure 4.20: Average boresight gain (top) and standard deviation (bottom) for a square quad ridge horn over varying flare angle and length at ridge exponential growth rates of (a) 1 mm⁻¹, (b) 51 mm⁻¹ and (c) 101 mm⁻¹. Note that the presented values are in dBi.

As seen, wider flare angles correspond to a greater aperture phase error and therefore lower gain and broader beamwidth, even though the aperture is electrically larger. For a given flare angle varying the length simultaneously varies the aperture size, which result in two major effects for which a balance must be sought. The first is that the low-frequency gain increases as a result of the larger aperture size. However, at higher frequencies there is a gain dip. The dip is caused by higher-order mode contamination
generated at the throat, that shifts lower in frequency and increases in severity as aperture size is increased. Therefore a careful trade must be sought that balances the high- and low-frequency gains. An abrupt transition between the first and second section ridge cause modal contamination to occur, which distorts the pattern from its desired shape. Specifically, the pattern develops a dip in gain at boresight. Besides constant boresight gain, this modular horn extension is able to be used for other far field parameters. The next parameter considered is constant E or H plane beamwidths. Contour plots for this is shown in Figures 4.21 and 4.22.

Figure 4.21: Average E plane 3 dB beamwidth (top) and standard deviation (bottom) for a square quad ridge horn over varying flare angle and length at ridge exponential growth rates of (a) 1 mm⁻¹, (b) 51 mm⁻¹ and (c) 101 mm⁻¹. Note that the presented values are in degrees.
Figure 4.22: Average H plane 3 dB beamwidth (top) and standard deviation (bottom) for a square quad ridge horn over varying flare angle and length at ridge exponential growth rates of (a) 1 mm\(^{-1}\), (b) 51 mm\(^{-1}\) and (c) 101 mm\(^{-1}\). Note that the presented values are in degrees.

As seen, there are certain values of flare angle, flare length and exponential taper rate which allows for a constant beamwidth in either the E or H plane with variation of less than 5 degrees. Note that constant beamwidth of less than 5\(^{\circ}\) is not achievable in both planes due to the modal content present at the square radiating aperture. Constant E plane beamwidth is in the middle right side of the provided contour plots, whereas constant H plane beamwidths are achievable in the top right corner. For both configurations, there is not a constant beamwidth greater than 30\(^{\circ}\) with less than 5\(^{\circ}\) of variation available, due to the large radiating aperture.
Figure 4.23: Fabricated (a) 13 dBi, (b) 15 dBi and (c) 18 dBi constant boresight gain horn extensions.

To show the validity of the provided contour plots, three different constant boresight gain horn extensions with variation less than ± 1 dB over the band are chosen for fabrication and testing. Specifically, 13 dBi, 15 dBi and 18 dBi constant gain horn profiles are chosen. The 13 dBi constant gain horn has a flare length of 1000 mils, flare angle of 30°
and exponential ridge taper of 76 mm$^{-1}$. The 15 dBi constant gain horn has a flare length of 1500 mils, flare angle of 25° and exponential ridge taper of 51 mm$^{-1}$. The 18 dBi constant gain horn has a flare length of 2000 mils, flare angle of 20° and exponential ridge taper of 26 mm$^{-1}$. Final fabricated models are shown in Figure 4.23. Each horn is fabricated through the 3D printing process detailed in Chapter 6, and then plated at Repliform, Inc. [127]. Note that the 15 dBi constant gain horn does not have full copper plating in the interior like the other two fabricated models. However, there is enough copper deposited on the interior for proper radiation as will be seen. Each horn extension is connected to the previously designed OMT with corrugated cone matching stub.

Measured results for the 13 dBi constant boresight gain horn are shown in Figures 4.24 – 4.26.

![Figure 4.24: Measured and simulated VSWR and boresight gain for the 13 dBi constant boresight gain horn extension.](image-url)
Figure 4.25: Measured (hashed) and simulated normalized radiation patterns for the 13 dBi constant boresight gain horn extension.

Figure 4.26: Measured (hashed) and simulated E and H plane 3 dB beamwidths for the 13 dBi constant boresight gain horn extension.
As seen, good agreement with simulations is obtained. Measured VSWR for both inputs is below 1.8 while the gain is $12.8 \pm 1$ dBi. Radiation patterns match well to simulated values. H plane radiation patterns are smooth, while there is rippling in the E plane due to the abrupt transition between the two horn interfaces. E plane beamwidth varies from 65° to 45°, while H plane beamwidth varies between 45° and 30°. Note that at lower frequencies the E plane beamwidth is larger than simulated values, which was also observed in small aperture only measurements. Discrepancies are mainly attributed to roughness from the 3D printing process, and misalignment of the secondary horn to the OMT output. While efforts are made to smooth the aperture walls prior to the plating process, there is still a roughness which the plating process accentuates. Results obtained for the 15 dBi constant gain horn are shown in Figures 4.27 – 4.29.

![Figure 4.29: Measured and simulated VSWR and boresight gain for the 15 dBi constant boresight gain horn extension.](image-url)
Figure 4.30: Measured (hashed) and simulated normalized radiation patterns for the 15 dBi constant boresight gain horn extension.

Figure 4.31: Measured (hashed) and simulated E and H plane 3 dB beamwidths for the 15 dBi constant boresight gain horn extension.
Measured VSWR for both inputs is below 1.8 while the gain is $14.5 \pm 1.25$ dBi. Radiation patterns match well to simulated values. As before, H plane radiation patterns are smooth, while there is rippling in the E plane due to the abrupt transition between the two horn interfaces. E plane beamwidth varies from 45° to 30°, while H plane beamwidth varies between 35° and 25°. The H plane beamwidth shows low variation over frequency for this specific configuration. Discrepancies are mainly attributed to roughness from the 3D printing process. Results obtained for the 18 dBi constant gain horn are shown in Figures 4.32 – 4.34.

![Figure 4.32](image)

Figure 4.32: Measured (hashed) and simulated VSWR and boresight gain for the 18 dBi constant boresight gain horn extension.
Figure 4.33: Measured and simulated normalized radiation patterns for the 18 dBi constant boresight gain horn extension.

Figure 4.34: Measured and simulated E and H plane 3 dB beamwidths for the 18 dBi constant boresight gain horn extension.
Measured VSWR for both inputs is below 1.85 while the gain is 17.5 ± 1.5 dBi. Radiation patterns match well to simulated values. H plane radiation patterns are smooth, while there is asymmetric rippling in the E plane due to the abrupt transition between the two horn interfaces. E plane beamwidth varies from 35° to 25°, while H plane beamwidth varies between 32° and 20°. Note that the E plane 3 dB beamwidth increases midband due to the introduction of higher order modes in the aperture.

4.4 Summary

An orthomode transducer based upon a custom double ridge cross section has been designed and fabricated for use in an 18 – 45 GHz towed decoy platform for dual polarized electronic attack. To provide for varying options of impedance match, isolation and power handling several different matching stub sections in the turnstile junction are shown. Each matching stub has an impedance match better than -10 dB over the band, with measured isolation greater than 35 dB. To provide greater flexibility in needed performance characteristics, a modular horn section is then designed and directly attached to the small radiating aperture of the OMT. This modular horn is able to be designed for performance criteria such as constant gain, beamwidth, etc. A 13, 15 and 18 dBi constant gain horn extensions are fabricated and measured, with results comparing favorably to simulations.
CHAPTER 5

Multiple Beam Generation for Electronic Attack

5.1 Introduction

As seen in Chapters 2 and 3, beamforming networks are used to feed antenna arrays and thus generate multiple beams. The basic building blocks of these networks, such as the 90° hybrid and Rotman lens, are implemented and designed differently depending on the transmission line used, manufacturing technology, center frequency, bandwidth of operation, and the desired coupling and isolation [128]. In single conductor waveguide structures, typically used for high power electronic attack applications, the design methodologies for hybrid couplers are typically either through the use of apertures [129] – [132] or branch guides [133] – [136].

Aperture based designs use Bethe’s small hole coupling theory to generate desired performance [137]. This can be done in either the broad wall or side wall of waveguide structures, wherein the latter are often limited in bandwidth [138]. Extensive studies have been conducted on the design of these couplers, focusing on parameters such as the slot shape, distances between slots and the slot dimension. For wider bandwidth, the same theory can be used with ridged waveguides. It is a relatively simple extension to use small hole aperture theory for single ridged waveguide. Double ridge waveguide couplers are typically based on the use of broadwall slots [139] – [140].

Rotman lens designs have been extended into the millimeter waves through new fabrication techniques such as substrate integrated waveguide (SIW) [141] and low
temperature co-fired ceramic (LTCC) [81]. However, losses in these transmission lines can be prohibitive for the applications discussed in this thesis. Rectangular waveguide based designs are often used for the Rotman lens beamformer [142]. However, these high power capable designs are inherently limited to a single modal bandwidth.

In this Chapter we demonstrate the design of beamformers based upon the double ridge waveguide topology introduced in Chapter 4. The first beamformer considered is a multi-section branch guide 90° hybrid. The designed hybrid contains 8 coupling sections with the same impedance at each shunt section. A Rotman lens beamformer using the designed double ridge waveguide topology is then investigated. By properly choosing Rotman lens design parameters, a lens with losses comparable to rectangular waveguide based designs can be achieved for high power electronic attack applications.

The chapter is organized as follows:

- **Section 5.2** discusses an 8 section branch guide 90° hybrid. The design procedure for this hybrid is discussed first, followed by fabrication using CNC machining. The fabricated hybrid is then measured and compared to simulations. It is seen that measured return loss and isolation are > 12 dB over an 18 – 45 GHz band, and measured phase difference between output ports is 90° ± 5°. Measured amplitude difference is 2.5 dB ± 1.25 dB, and is larger than expected due to burrs from the CNC machining process.

- **Section 5.3** overviews a Rotman lens beamformer based on double ridge waveguide. A parameter study is undertaken to determine effects on impedance and loss in the lens region. Based on these studies, a topology is then chosen with VSWR < 1.5 for all input ports and insertion loss < 2 dB. The lens is fabricated using split block CNC
machining and measured. Results show consistent H plane beam location over the desired bandwidth, and VSWR < 2.5.

5.2 Branch Guide Coupler

As discussed in the introduction, waveguide couplers commonly use either aperture or branch guide topologies. In the millimeter waves, required double ridge waveguide size constraints can make the fabrication of apertures difficult using traditional techniques. Therefore, a branch guide topology is chosen for the 90° hybrid.

Figure 5.1: Examples of waveguide cross sections having identical $Z_{PI}$ versus frequency [142].

In the design of this hybrid, it is key to note that two waveguide modes with the same cutoff frequency have characteristic impedance ratios that are independent of
frequency. This observation has been verified for hollow metallic waveguides of many different cross-sections. This observation appears to be valid when comparing a single definition of $Z_0$, such as the power-current definition $Z_{PI}$, the power-voltage definition $Z_{PV}$, or the voltage-current definition $Z_{VI}$. For $Z_{PV}$ and $Z_{VI}$ the comparisons must be made for a given integration contour. As a consequence, many different cross-sections can realize a given characteristic impedance by careful design. Example cross sections with the same impedance are shown in Figure 5.1. Another consequence is that for a given cross section, a second cross section can be constructed that has identical phase characteristic but different impedance. This has been exploited to create waveguides with the necessary branch impedances and thus realize wideband couplers [142].

The design challenge herein arises from the selection of the branch guide coupling cross section. Ideally, the coupling sections have impedance that is proportional to the main guide impedance, independent of frequency, while the phase characteristic of the branch guides is identical to that of the main guide. Also, the single-mode bandwidth of the coupling section must prevent spurious resonances. Lastly, the mode field structure should match as closely as possible to the mode field in the main guide. Deforming a double-ridge waveguide, excited in the $TE_{10}$ mode, such that the new boundary is tangential everywhere to the unperturbed magnetic field produces a waveguide cross section with the same cutoff frequency and frequency-independent characteristic impedance ratios as the original waveguide. An advantage of cross-sections using this technique is that the cross-section widths are equal to those used in the original guide. This results in a coupling characteristic with improved flatness versus frequency. Figure 5.2 shows the development of the coupling cross-section.
Although the couplers exhibit excellent broad-band performance in simulation, there is a key aspect to the couplers that makes fabrication a challenge. As a result of using the above design process, the branch guide coupling sections have narrow ridge gaps that make them challenging to fabricate. An eight section design is chosen to strike a balance between desired electrical performance and the ability to fabricate each section. The respective gaps used in the eight section design are 6 mils (152 μm) at the center and 23 mils (584 μm) at the edge. Note that larger gaps can be used at the expense of worse amplitude misbalance over the bandwidth. Coupling sections are separated by 94.5 mils (2.4 mm), while the through section double ridge waveguides are separated by 135.6 mils (3.44 mils), or 16.3 mils (414 μm) between bottom and top waveguide walls. To facilitate measurement, a transition section is designed to route the double ridge lines far enough away to attach WRD 180 coaxial adapters. The final computer simulated model is shown in Figure 5.3, a representative CNC machined piece in Figure 5.4 and measurement setup in Figure 5.5.
Figure 5.3: Computer generated model of the eight section 90° hybrid.

Figure 5.4: Machined section of the 90° hybrid. Note that there are 9 total pieces that are affixed together using alignment pins and screws.
An Agilent N8363B VNA is used for S-parameter measurements, and obtained reflection coefficient and isolation are compared to simulated values in Figure 5.6. Note that this VNA is limited to a maximum frequency of 40 GHz. As seen, measured reflection coefficient and isolation are less than -12 dB over the 18 – 40 GHz range. Measured and simulated amplitude misbalance and phase difference are shown in Figure 5.7. The measured phase difference between the output ports is 90° ± 5°. Note that increased amplitude misbalance is seen as compared to simulated values. To see this behavior, the coupled and through port magnitudes are shown in Figure 5.8. Increased loss of about 2 dB is seen for the coupled port as compared to simulated values, while the through port has 4 dB more loss. This leads to increased amplitude misbalance of the hybrid.

Careful investigation of the machined pieces (see Figure 5.9) gives an explanation for the electrical performance. Burrs resulting from the CNC machining process are present in the waveguide structure. Burrs are deformed material that remains on a piece
after machining. It is often in the form of a rough strip of metal at the edge of the piece adjacent to the machined surface. This is a manufacturing phenomenon that is a common problem in many industries [144]. The removal of these burrs is typically done through abrasives or finishing tools, but can yield unpredictable and inconsistent results. Other fabrication methods such as wire EDM or laser cutting may be used with better fabrication success, but they were not actively pursued at this time.

Figure 5.6: Measured reflection coefficient and isolation of the 90° hybrid.

Figure 5.7: Measured amplitude misbalance and phase difference of the 90° hybrid.
Figure 5.8: Measured coupling and through port magnitudes of the 90° hybrid.

Figure 5.9: Close up view of a single machined piece of the 90° hybrid. Burrs resulting from the CNC process affect the electrical performance.

5.3 Double Ridge Rotman Lens

Planar microwave lenses like the Rotman lens [74] are good candidates for many millimeter wave applications including communications, automotive radar and biomedical
imaging because of their ability to form multiple beams at wide angles. The Rotman lens exhibits a theoretically large bandwidth due to its true time delay (TTD) beamforming nature. They have been studied for several decades with many devices demonstrated in various technologies including the most common microstrip and rectangular waveguide-based. However, lenses built using rectangular waveguides suffer from higher order modes, which limit their bandwidth to less than an octave, because of the parallel plate geometry of the lens. Microstrip lenses can be built at these frequencies, but they often have high loss. To alleviate the above discussed issues, a double ridged waveguide Rotman lens with single modal operation over an 18 – 45 GHz band is designed.

As discussed in Chapter 3, careful design of the double ridge waveguide to parallel plate transition is crucial. Ideally, the ridges have an exponential taper into the top and bottom walls along with an expansion of the overall waveguide dimensions, as in [87]. For this design, the waveguide width exponentially flares from the initial 265 mils (6.73 mm) to 500 mils (12.7 mm), while the waveguide height flares from 119 to 150 mils (3.02 to 3.81 mm). The ridge extends into the parallel plate cavity 200 mils (5.08 mm). The final length and width of the transition is 800 mils (20.32 mm) × 500 mils (12.7 mm), respectively, and is shown in Figure 5.10. The simulated performance of the transition is presented in Figure 5.11. As seen, the transition has a reflection coefficient less than -25 dB over the 18 GHz to 45 GHz bandwidth, with insertion loss below 0.05 dB.

To demonstrate the utility of the developed double ridge to parallel plate transition for a millimeter wave Rotman lens, a 3 input 5 output port device operating over an 18 – 45 GHz band is desired. To generate an appropriate geometry for electronic attack applications, special care in choosing geometrical parameters is needed. It is well known [88] that besides array size, the critical design parameters for a Rotman lens are the focal length and focal angle.
Therefore, with an array element separation of 6.5 mm (chosen for the provided antenna array), 5 beam ports and center frequency of 30 GHz, the focal length and focal angle are varied to find a lens topology with low insertion loss and good impedance match
at the 3 focal points on the beam contour. Here, insertion loss is computed as the sum of residual powers that are available at the array ports. For this design, sidewalls terminated with absorber are used due to readily available broadband absorber in the 18 – 45 GHz frequency range. It is understood, due to reported power handling capabilities of commercially available absorber, an alternative solution is needed (such as the dummy ports used in Chapter 3). The focal angle is the parameter studied first. On the beam contour, the focal angle determines where (theoretically) zero phase error between array ports occurs besides the center of the contour. The focal length chosen is $5\lambda_{30\text{ GHz}}$ (5 cm), and the focal angle is varied between 15° and 45°. All simulations are performed through full wave simulations. Considered designs are shown in Figure 5.12, and simulated VSWR and insertion loss in Figure 5.13.

![Figure 5.12: Top views of the Rotman lens when the focal angle is changed.](image)
Figure 5.13: Performance of the Rotman lens when the focal angle is changed. Shown are the (a) VSWR and (b) insertion loss for both inner and outer beam ports.

As seen, wider focal angles correspond to increased insertion loss for outer beam ports, in addition to deteriorated VSWR. This is due to lesser power delivered to the array ports, and more power delivered to the dummy load. Insertion loss and VSWR for the inner beam port are relatively unchanged for the generated designs. To keep insertion loss as low as possible, the focal angle is chosen to be 15°. The focal length is then varied between $3\lambda_{30\text{ GHz}}$ (3 cm) and $7\lambda_{30\text{ GHz}}$ (7 cm). Created designs are shown in Figure 5.14, and simulated VSWR and insertion loss in Figure 5.15.
Figure 5.14: Top views of the Rotman lens when the focal length is changed.

Figure 5.15: Performance of the Rotman lens when the focal angle is changed. Shown are the (a) VSWR and (b) insertion loss for both inner and outer beam ports.
As seen, a shorter focal length produces slightly better insertion loss and VSWR results for both inner and outer beam ports. As before, this is because the array ports receive more of the beam power from the parallel plate region. A consideration to make when choosing the focal length is that a shorter focal length lens can make the addition of additional beam and array ports difficult to integrate.

From this design study, a Rotman lens configuration with low insertion loss is chosen. A 3 beam port 5 array port lens with a focal length of $4.5\lambda_{30 \text{GHz}}$ (4.5 cm), focal angle of 12.5°, 6.5 mm spacing between array elements and focal ratio of $g = 1.0238$ is designed. Sidewalls are chosen to be lined with HR-25 broadband absorber from Laird Technologies [145]. Note that further studies on the power handling capability of the absorber are needed before this lens could be used for high power applications. A transition to dummy port sidewall, as presented in Chapter 3, is a viable option if the absorber sidewall cannot handle the desired power levels. A view of the lens only design is shown in Figure 5.16. The lens has a width of 3.225 in. (8.19 cm), length of 3.4 in. (8.64 cm) and thickness of 0.15 in. (0.38 cm). Simulated VSWR, coupling and insertion loss are shown in Figures 5.17 – 5.19.

As seen, the simulated VSWR and coupling are below 2.25 and 10 dB, respectively, over the 18 – 45 GHz band for all input ports. Note that there is more coupling between the outer ports than the adjacent inner beam port due to outer beam ports being symmetrical, so any power reflected from the array contour is aligned directly with the opposite beam port. The primary design consideration for this Rotman lens is the insertion loss. Loss of less than 2 dB is seen over the bandwidth for both inner and outer beam ports. This is an improvement as compared to the double ridge waveguide lens presented in Chapter 3.
Figure 5.16: Top views of the designed Rotman lens.

Figure 5.17: VSWR for both inner and outer beam ports for the designed Rotman lens.
Properly phased lines and a double ridge waveguide antenna array are integrated with the lens to measure performance. The final lens is 8.25 in. (20.96 cm) × 5 in. (12.7 cm) × 1 in. (2.54 cm), and is constructed through split block machining. The fabricated pieces and construction process are shown in Figure 5.20. Notice that the two pieces are mirror images of one another. The phased line section is 2.78 in. (7.06 cm) long and the array is
2.3 in. (5.84 cm) long. Note that these two sections make up over 60% of the total length, due to the size of the double ridge waveguide cross section in the phased line region. The sidewall absorber section has dimensions of 1 in. (2.54 cm) × 1.2 in. (3.05 cm). The final aperture size is 1.29 in. (3.28 cm) × 0.5 in. (1.27 cm). Measured S parameters are shown in Figures 5.21 and 5.22.

Figure 5.20: Fabrication of the 18 – 45 GHz low insertion loss Rotman lens. (a) shows one half of the lens and (b) shows the other half of the lens with absorber in the sidewall cavity. (c) the two pieces are then connected together through alignment pins and screws.

An Agilent N8363B VNA is used for S-parameter measurements, and obtained reflection coefficient and isolation are compared to simulated values in Figure 5.21, and coupling between beam ports in Figure 5.22. Note that this VNA is limited to a maximum frequency of 40 GHz. As seen, measured VSWR is less than 2.5 over the band. Measured
results are worse than simulated values due to misalignment of the WRD-180 coaxial adapters used for measurement. This can be mitigated in the future with more careful fabrication of the lens pieces.

Figure 5.21: VSWR of both inner and outer beams of the fabricated Rotman lens.

Figure 5.22: Coupling between the inner and outer beam ports of the fabricated Rotman lens.
Measured coupling between outer and inner beam ports is similar to simulated values, and is less than -12 dB over the band. To show that the desired far field performance of the designed lens is still achieved even with a higher VSWR, normalized gain plots for both inner and outer beam ports is shown in Figure 5.23.

Figure 5.23: Simulated and measured normalized gain plots for (a) inner beam port and (b) outer beam ports.
As seen, measured gain plots for both inner and outer beams are similar to their simulated values. There is consistent beam location for the outer beam port as expected for the Rotman lens beamformer. While there is increased lobing at the lower frequencies due to the performance of the attached antenna array, the purpose of this lens was to show that a low insertion loss lens can be designed and fabricated.

5.4 Summary

This Chapter has shown the development of wideband beamformers based upon the custom double ridge waveguide cross section introduced in Chapter 4. An 8 section branch guide 90° hybrid is first presented. Measured results show a return loss and isolation greater than 12 dB over the desired 18 – 45 GHz bandwidth, with phase difference of 90° ± 5°. While the amplitude misbalance is up to 4 dB between the ports, a close examination of the fabricated pieces has shown that burrs from the CNC fabrication process are the issue. This can be resolved through better fabrication processes in the future. A Rotman lens topology is then designed and fabricated. A parameter study is undertaken to see how the focal angle and focal length affect impedance and insertion loss in the lens. A 3 beam, 5 array element Rotman lens with focal angle of ± 12.5° is chosen and fabricated using split block CNC machining. Measured results confirm the generation of appropriate beams, with coupling between beam ports < -12 dB over the band and VSWR < 2.5.
CHAPTER 6

Prototyping of Electronic Warfare Front-End Subsystems

6.1 Introduction

Emerging technologies and fabrication processes including 3D microfabrication enable the development of passive and active components and subsystems for various applications into the millimeter and sub-millimeter waves [146]. 3D microfabrication allows for a high degree of integration, tight packaging, exploitation of the third dimension, reduced form factors and footprints as well as low loss and dispersion. Of particular interest are rectangular coaxial transmission line-based components and subsystems built using different approaches including EFAB [147], SU-8 plating [148] and PolyStrata [33]. While manufacturing quality, precision and yield have much improved, there is still a substantial cost and design complexity involved in using these fabrication processes.

RF and millimeter wave systems can be expensive to design and develop and often require substantial labor commitments to manufacture. As these costs and design complexity increase, accurate simulations and prototype concept analysis have emerged as a valuable design aid. Although computer simulated models can replicate system performance with good results [149], the simulation of entire systems can be difficult to setup. By physically constructing prototypes, the designer is able to investigate different conceptual approaches for their strong points and shortcomings. In addition, it can help to prevent over-specified system designs and allow a favorable trade-off between selected components from a cost and technical point of view.
This chapter investigates PCB prototyping and 3D printing to prototype RF and millimeter wave antennas and subsystems. First, a low-cost, fast turnaround prototyping approach for the above discussed 3D microfabrication techniques based on stacked printed circuit boards (PCB) is proposed [150]. This approach mimics surface microfabrication processes in creating microwave components and subsystems layer by layer. The resultant process is able to produce good quality, very low-cost components and subsystems, as demonstrated by a fabricated 4 to 8 GHz two element tapered slot antenna array with integrated 90° hybrid. Wide impedance bandwidth with nominal gain around 10 dBi, 3 dB beamwidth between 30° and 45°, consistent beam pointing with predictable nulls, and excellent agreement with simulations all testify to the merit of the proposed approach. Good performances of the stand-alone 90° hybrid and two element antenna array over the same or even wider bandwidth are also demonstrated. Second, a 3D printed millimeter wave horn antenna is designed and compared to a conventionally machined antenna and simulations. Measured VSWR below 2, gain between 5 dBi and 10 dBi, and wide E and H plane beamwidths are observed over an 18 – 40 GHz range. Good correlation with the EDM horn and simulations confirms the ability of 3D printing to prototype RF and millimeter wave antennas and subsystems.

This chapter is organized as follows:

- Section 6.2 discusses the use of stacked PCB boards for prototyping 3D-micromachined front-ends. Design and fabrication of a 4 – 18 GHz standalone 2 element DF array and an integrated 4 – 8 GHz amplitude only DF array are used to demonstrate the utility of the proposed technique.

- Section 6.3 details the 3D printing process for prototyping a small aperture horn and compares obtained measurements with a professionally built antenna. Favorable
results and physical/cost properties indicate that 3D printing may be used as a good approach for prototyping appropriate millimeter wave devices.

6.2 PCB Prototyping

Because of its ultra-low loss, dispersion, and cross-talk, as well as tight packaging and ability to reduce size by utilizing the third dimension, an air filled rectangular coaxial line in millimeter waves has received increased recent attention and may be considered as an alternative to traditional transmission line technologies such as microstrip or stripline. Smaller form factors and wider bandwidths with regards to rectangular waveguide or substrate integrated waveguide can also be achieved. While many passive components such as hybrids and filters have been successfully demonstrated with these technologies, antenna fabrication, design and direct integration with beamforming networks presents additional challenges. To mitigate these challenges, low-cost, fast turnaround prototyping approach for the above discussed 3D microfabrication techniques based on stacked printed circuit boards (PCB) is proposed. This approach mimics surface microfabrication processes in creating microwave components and subsystems layer by layer.

Components and subsystems built by 3D surface micromachining processes share some common fabrication steps including: (i) sacrificial material is deposited on a substrate; (ii) metal layer is deposited over this pattern; (iii) planarization process forms a single layer composed of the two different materials; (iv) sequence (i) – (iii) is repeated over many layers; (v) the sacrificial material is then etched away, leaving only the metal frame. To allow for the removal of sacrificial material, it is common to have release holes in outside walls. Proper design of these holes is critical, as the holes must be large enough to allow
the sacrificial layer to be removed, while preventing radiation and maintaining structural stability. There must also be some kind of mechanical support for the inner conductor of the coaxial line. As enabled by a specific microfabrication process, this can come from dielectric membranes, dielectric stubs, metal posts, or the metal stubs as used in this work.

A major drawback of these fabrication processes is they are time-consuming (both design and fabrication) and often expensive. Thus, to avoid having faulty components and subsystems produced, it is desirable to first have a prototype test bed. To fill this need, we use a PCB configuration mimicking the typical layering sequence used in surface micromachining. By copper plating the slotted sidewalls on the PCB, a low-loss, non-dispersive, high-isolation 3D transmission line can be constructed with the stacked procedure. This is a common procedure for practically all PCB foundries, with quick fabrication turnaround and low cost. As an illustration of the basic transmission line that can be built by stacking PCB board, a five board layout of an air-filled rectangular coax line and traditional rectangular waveguide is presented in Figure 6.1. Note this is the minimum number of layers required to construct an air-filled rectangular coaxial line. The impedance of the rectangular coax line and rectangular waveguide can be controlled by the height of the stacked boards and the widths of the inner conductor and air gap. The PCB layers can be held together by either epoxy or screws to provide an electrical connection between layers. Note that air gaps between layers can occur with the use of either epoxy or screws. After careful study, it is seen that the effect of these air gaps is small compared to PCB manufacturing tolerances, as presented later. This is not surprising as they are mostly parallel to the current flow in the outer conductor. The number of boards and height of each board can be varied as needed to replicate desired features of micromachining techniques and achieve lines, components and systems with desired electrical performance.
A limiting factor for this approach is that of the circuit board manufacturing process, specifically the available route bit size. In prototyping at microwave frequencies, a smaller route bit size is needed to ensure single mode operation, and thus accurate prototyping for the stacked PCB approach. The available panel size, substrate type, copper thickness, minimum routed slot width, roughness in plated sidewalls, and manufacturing tolerances are also important to consider during the design of the prototyping subsystem.

### 6.2.1 Two Element Vivaldi Array Development

To demonstrate the use of the PCB stacking for prototyping millimeter wave systems and systems in their own outright, a two element Vivaldi antenna array is designed for use as a baseline element of amplitude/phase and amplitude only DF. A basic coplanar Vivaldi antenna is shown in Figure 6.2.
The antenna is fed perpendicular to the open end. The minimum operating frequency is determined by the length of $W_{\text{slot}}$ that is $\approx \lambda/2$. The maximum operating frequency is set by either where the length of $W_{\text{gap}}$ is $\approx \lambda/2$, or the length of the open end is $\lambda/4$, whatever frequency is higher. The exponential taper of the Vivaldi geometry is obtained from

$$y = c_1 e^{R_g x} + c_2 \quad (6.1)$$

where

$$c_1 = \frac{1}{2} \frac{W_{\text{slot}} - W_{\text{gap}}}{\exp[R_g L_{\text{slot}}] - \exp[R_g L_0]} \quad (6.2)$$

$$c_2 = \frac{1}{2} \frac{W_{\text{gap}} \exp[R_g L_{\text{slot}}] - W_{\text{slot}} \exp[R_g L_0]}{\exp[R_g L_{\text{slot}}] - \exp[R_g L_0]} \quad (6.3)$$

$c_1$ and $c_2$ are constants, and $R_g$ is the opening rate of the exponential flare. A smaller opening rate yields better performance in the lower frequency band at the expense of the higher frequency band. A small taper length affects the front-to-back ratio of the radiation patterns at the lowest frequency, whereas a longer taper length provides better performance up to fabrication limits [151]. When two or more Vivaldi antennas are combined to form an array, the mutual coupling between the elements decreases the lowest frequency of operation [61].
An all metal-design of the antenna array is considered for this study. The initial configuration based upon an $1 \times 8$ array [61] and its performance at $f/f_0 = 2.5$ are shown in Figure 6.3. The amplitude/phase DF system is considered primarily to optimize the antenna array using the method of moments code FEKO [66]. As seen in Figure 6.3, the radiation pattern of this two element array is not symmetrical around the boresight direction. In addition, the difference mode VSWR of the array is not $< 3:1$ until $f/f_0 > 2$, which affects the overall size of the antenna array. To improve the performance of the antenna, the geometry was made symmetrical as shown in Figure 6.4. By making the antenna symmetrical, the radiation pattern of the difference mode becomes smoother. In addition, there is a slight improvement in the respective VSWR of the difference mode.
Figure 6.3: Initial two element antenna array design for an amplitude/phase DF system based off an 8x8 array design from [9] (a), computed VSWRs (b), and radiation patterns at $f/f_0 = 2.5$ (c) for the sum and difference modes.
Figure 6.4: (a) Symmetric two element antenna array compared to the initial design from Figure 6.3, and (b) computed VSWRs, and (c) radiation patterns at $f/f_0 = 2.5$ for the sum and difference modes.

To further improve the antenna array match, the exponential opening rate of the flare was increased from 0.0006 to 0.1 after a parametric study. Figure 6.5 shows the simulated comparison between the nearly linear taper from the original design and the exponential taper (Vivaldi). As expected, the VSWR of the sum and difference modes both improve with the redesign, with the difference mode VSWR $< 3:1$ from $f/f_0 > 1.5$. However, the radiation patterns for the sum mode are not as wide as in the linear taper case, which can affect the DFF range. In addition, the difference pattern rippling becomes more pronounced. Because the DF function for the amplitude/phase system is determined by the first null of the sum mode, the rippling in the difference pattern will not severely affect the DFF.
Figure 6.5: (a) Two element array with increased exponential rate, (b) VSWR and (c) pattern comparisons at $f/f_0 = 2.5$ with the system in Figure 6.3.

To reduce the size of the two element array even further, corrugations were introduced onto either side of the antenna as demonstrated for a single Vivaldi antenna in
A longer corrugation length improves the E plane beamwidth of a single antenna element at the expense of the H plane beamwidth. The same effect is observed by widening the corrugations [153]. It is important to note that corrugations increase the side lobe level and moved the sum pattern null to lower elevation angles. While the former is important for DF jammers, the latter will produce a narrower field of view for the desired DF system. Corrugations optimized using a genetic algorithm were added to the two element antenna array, with the design and obtained results shown in Figure 6.6.
Corrugations lower the difference mode VSWR to below 3:1 at \( f/f_0 > 1 \). The sum mode is relatively unchanged at lower values of \( f/f_0 \), but the VSWR rippling is reduced at higher frequencies. The beamwidth of both the sum and difference modes is improved with the corugations. In addition, the length of the antenna array is reduced when compared to previous designs. While the array design was used to mainly benefit the amplitude/phase DF technique, there is also a benefit to the squinted beams produced in the amplitude only DF technique. First, the effect of the growth rate on a two element array with connected 90° coupler is shown in Figure 6.7. As seen, there is a specific growth rate where the turn on frequency is lowest as compared to the width of the two element antenna array, which allows for a smaller array. However, at this specific growth rate, the gain is near its minimum value, and as expected the 3 dB beamwidth is near its maximum. To increase the gain of the antenna array at this growth rate, corrugations are introduced to the outer edges of the two element antenna array. The addition of corrugations leads to increased gain by 2.5 dB, and decreased 3 dB beamwidth from 65° to 45° for the same sized array.
Figure 6.7: (a) The effect of the antenna growth rate on the turn on frequency, gain and 3 dB beamwidth of the two element antenna array, and (b) the effect of corrugations on the radiation patterns. The normalized wavelength is equal to the width of the two element antenna array.
6.2.2 Two Element Array Measurements

The two element antenna array from previous section was fabricated using a local PCB foundry. To achieve the correct heights (and therefore impedances) for 4 – 18 GHz operation, a five board layout using 0.039” FR4 board and 1 oz. copper thickness is designed. The layers are separated into two separate 6” x 9” layouts, with the center coaxial line layer repeated twice. The fabricated boards are shown in Figure 6.8, and are scored for easy removal of individual layers. The layers and completed devices are shown in Figure 6.9 with Southwest Microwave SMA end launch connectors [155].

Figure 6.8: Photographs of the fabricated PCB boards for the two element tapered slot antenna array.
The antenna array is directly connected to wideband COTS couplers for measurements. The chosen frequency range is 4 – 18 GHz, with the 180° coupler from RF Lambda [156] and the 90° coupler from Pulsar Microwave [157]. An Agilent N5246A PNA is used for impedance measurement. Obtained VSWRs are compared to simulated data and the results are shown in Figure 6.10.

Figure 6.9: Fabricated PCB boards for the two element tapered slot antenna array. Shown are the (a) individual layers and the (b) complete array.

As seen, the VSWRs of the sum and difference modes as well as those obtained using the 90° hybrid compare well to simulations. Both modes of an amplitude/phase front end are matched to VSWR < 2.5:1 throughout the entire band (note that VSWR < 2:1 for most of the range). The output ports of an amplitude only front end have VSWR < 2.5:1 from 4 – 18 GHz. Note that the COTS 90° coupler has a lower level of misbalance than the 180°
coupler, which will lead to the lesser deviations of far-field data from the ideal beamforming networks.

Figure 6.10: Simulated and measured VSWRs for both (a) amplitude/phase and (b) amplitude only DF systems.

The radiation patterns are measured in a University of Colorado combined spherical near field/far field anechoic chamber, as shown in Figure 6.11. The arrays are connected to their respective beamformer through a standoff connector. E-plane radiation patterns are
collected over a 4-18 GHz range and then normalized to the largest values. Data at three different frequency points are shown in Figure 6.12. As seen, all three radiation patterns show expected directive performance consistent with the design of a Vivaldi antenna array. Using a 180° coupler, the sum mode is oriented along boresight for all frequencies. The first null of the sum mode, which gives the unambiguous range of the amplitude/phase DF system, starts out at approximately ± 65°, and shifts inwards as frequency increases. This is expected because of side lobes that appear at higher frequencies. Side lobe levels are at least 10 dB below the boresight maximum value for the sum mode. The difference mode maximum is ± 40° at 4 GHz, and decreases to ± 20° at higher frequencies.

Figure 6.11: Experimental setup used to measure the normalized radiation patterns for both DF systems.
Figure 6.12: Normalized simulated and experimental radiation patterns for a DF system with a two element Vivaldi antenna array. Patterns are shown for amplitude/phase and amplitude only systems.
Figure 6.13: Normalized simulated and experimental co and cross polarized gains for amplitude/phase DF system with a two element Vivaldi antenna array.

Figure 6.14: Normalized simulated and experimental co and cross polarized gains for amplitude only DF system with a two element Vivaldi antenna array.

Far field parameters are also obtained for both DF systems, and are shown in Figures 6.13 – 6.15. As seen, for the amplitude/phase DF system, the sum mode gain varies between 9 and 11 dBi, while the difference mode varies between 4 and 8 dBi. Sum mode cross polarization is at least 25 dB down, while the difference mode are 20 dB down over the band. For the amplitude only DF system, the far field gain varies from 8 to 11 dBi. Measured cross pol levels are at least 20 dB down over the band. Increased levels of cross...
polarization are attributed to scattering from mounting devices. 3 dB beamwidths vary from $55^\circ$ to $27^\circ$ for the sum mode, $50^\circ$ to $18^\circ$ for the difference mode and $45^\circ$ to $25^\circ$ for the squinted modes produced by amplitude only direction finding. For all far field parameters, measurements match well to simulations.

![Figure 6.15: Normalized simulated and measured E-plane 3 dB beamwidths for both amplitude only and amplitude/phase DF systems with a two element Vivaldi antenna array.](image)

When a 90° coupler is used, squinted beams off boresight are produced as desired for amplitude only DF systems. The first null on the opposite side of maximum determines the unambiguous range of the amplitude only DF system. The first null is at $\pm 45^\circ$ at 4 GHz, and decreases to $\pm 15^\circ$ at 18 GHz. The side lobe levels are also higher, and their level increases with frequencies. At 18 GHz, they are only 7.5 dB below the maximum value of the radiation pattern. This represents a challenge for these types of systems due to the susceptibility to the side lobe jamming technique. The unambiguous range for the amplitude/phase and amplitude only systems was also calculated and results are shown in Figure 6.16. The field of view (FOV) is shown to vary from $130^\circ$ to $50^\circ$ in simulations for the amplitude/phase system, and from $90^\circ$ to $30^\circ$ for the amplitude only system. In both DF
systems, the experimental FOV is below the simulated values at all frequencies. This is because the simulated radiation patterns are smoother than measurements. It is important, however, to note that the simulated models do not take into account the mounting plates and other auxiliaries used to attach the array to the mast. It is highly likely that these are sources of additional rippling in radiation patterns that directly translates to rapid changes in the FOV. That said, the overall trend of FOV vs. frequency is maintained for both DF front ends.

![Field of View E Plane Amplitude/Phase](a) ![Field of View E Plane Amplitude Only](b)

Figure 6.16: Simulated and measured FOV for (a) amplitude/phase and (b) amplitude only DF systems.

### 6.2.3 Highly-Integrated DF Front-End System

To further demonstrate the utility of the PCB prototyping approach, a full amplitude only front end system is designed by integrating a 90° hybrid with the previously designed two element antenna array. The former is designed to work from 4 to 18 GHz, while the latter operates over a 4 to 8 GHz range, and is thus (for this specific configuration) limiting the bandwidth of the subsystem. A 90° hybrid is a $2 \times 2$ Butler matrix feed for two independent beams of a two element antenna array. It is also a basic building block of
more complex analog beamforming networks and corresponding multi-element arrays. A branchline 90° hybrid configuration is chosen since it does not require displacement current couplings and also because it is inherently compatible with the proposed approach. Since a traditional single section branchline 90° hybrid is narrowband, a multi-section topology from [159] is adopted. As each additional section is introduced, it is seen that the impedance of the outermost shunt branches increases, while the impedance of the series branch decreases. This can cause unreasonably high and low impedances difficult to manufacture in any process, including the approach discussed herein. Thus, a three section branchline coupler able to work over an octave bandwidth is chosen. To increase the impedance bandwidth of the hybrid, a matching section with shunted stub is introduced into each port [160]. The shunted stubs also provide mechanical support for the inner conductor, eliminating the need for mechanical supports such as dielectric straps. A circuit diagram is shown in Figure 6.17.

![Circuit diagram for an octave bandwidth 90° multi-section branchline hybrid designed for integration with a two element antenna array.](image)

Figure 6.17: Circuit diagram for an octave bandwidth 90° multi-section branchline hybrid designed for integration with a two element antenna array.
Because of the limited range of impedances available, circuit simulations are used to tune the impedances of the multi-section 90° hybrid. As seen in Table 6.1, the impedance of the outermost branchline shunt sections is 107 Ω, innermost branchline shunt section is 51 Ω, while the branchline series sections are 35 Ω. The matching line and matching shunt stub sections are 84 Ω and 70 Ω, respectively. The initial electrical length for each section of the line is 90°, except for the matching line length, which is 180° at the center frequency. To account for parasitics at line junctions, the final length tuning of each section is performed in HFSS and obtained results are also shown in Table 6.1. The final fabricated device is shown in Figure 6.18.

Table 6.1: Line lengths and impedance of the 90° hybrid from Figure 6.17.

<table>
<thead>
<tr>
<th>Line Section Name {Circuit Diagram Name}</th>
<th>Impedance [Ω]</th>
<th>Electrical Length [°] at Center Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>Matching Shunt {Shunt}</td>
<td>84</td>
<td>100.8</td>
</tr>
<tr>
<td>Matching Line {Line}</td>
<td>70</td>
<td>182.4</td>
</tr>
<tr>
<td>Branchline Shunt Outside {1}</td>
<td>107</td>
<td>90</td>
</tr>
<tr>
<td>Branchline Shunt Inside {2}</td>
<td>51</td>
<td>90</td>
</tr>
<tr>
<td>Branchline Series {3}</td>
<td>35</td>
<td>109.8</td>
</tr>
</tbody>
</table>

Figure 6.18: The photograph of the fabricated and fully assembled PCB stacked individual 4-8 GHz 90° hybrid.
An Agilent N5246A PNA is used for S-parameter measurements, and comparison with simulated results for a 90° hybrid is shown in Figure 6.19. As seen, the individual hybrid has a measured amplitude misbalance less than ± 2 dB, phase misbalance less than ± 7°, and reflection coefficient and port to port coupling less than -10 dB over the 4 to 8 GHz band. Note that though a lower level of port to port coupling is typically desired, obtained values are sufficient to assess the prototyping ability of the stacked-PCB approach.
Differences between the measurements and simulations are primarily attributed to the manufacturing tolerances of the PCB fabrication process.

The designed individual components as well as the complete subsystem are fabricated by standard circuit board manufacturing using seven FR4 boards with two different thicknesses. This number of boards and their thickness are necessary to achieve the needed impedances within the hybrid and antenna array. Layers 1, 4 and 7 are 0.508 mm thick, while layers 2, 3, 5 and 6 are 0.762 mm thick. Copper thickness of 1 oz. is used for both the horizontal and vertical (plated) sides of the lines and antenna array. The selected manufacturer allows for the minimum air gap width in the inner conductor layer of at least 0.762 mm and minimum routed slot width of 0.508 mm, which was implemented in the design. To assemble the device, 3.175 mm dowel pins are inserted to ensure proper layer alignment. The layers are then stacked to assemble the full component or subsystem. Once the final layer has been put on, screws are used to hold down the layers tightly. By using screws instead of epoxy, individual layers can be replaced with ease. Note that no attempts to treat the slotted walls of the individual components or subsystem are made.

The fabricated FR4 layers and construction process of the designed subsystem are shown in Figure 6.20. Nylon screws and nuts are used in the assembly of the individual antenna array and integrated antenna array subsystem to avoid undesired effects of metallic screws and nuts near the radiating slots. To complete the component and subsystem assembly, end-launch SMA connectors from Southwest Microwave are integrated with each coaxial launch.
Figure 6.20: (a) Computer generated illustration of the fabricated stack up and its assembly process. The assembly process and blown up view of the inner conductor layer showing the (b) shorting stubs and antenna connection, along with (c) the complete device.
To test component level prototyping using the PCB stacking process, the hybrid and two element tapered slot antenna array are directly integrated and fabricated, as shown above. Measured and simulated reflection coefficient and port to port coupling are shown in Figure 6.21. As seen, reflection coefficient and port to port coupling less than -10 dB are achieved throughout the octave bandwidth.

![Figure 6.21: Measured and simulated reflection coefficient and coupling between antenna ports for the complete two element antenna array.](image)

Measured and simulated E plane radiation patterns over the entire bandwidth are shown in Figure 6.22. As expected, beams squinted off boresight due to the integrated 90° hybrid are seen in both measurement and simulation with good correlation. The two beams are symmetric across the H-plane with null and side-lobe locations well predicted by full-wave simulations. Differences in null depth are attributed to increased phase misbalance in the fabricated system. Measured and simulated realized gain and 3 dB beamwidth are shown in Figure 6.23.
Figure 6.22: Measured and simulated normalized E plane radiation patterns for the complete two element antenna array subsystem.

As seen, realized gain between 9 and 11 dBi over the full bandwidth is measured. Differences between measured and simulated realized gains are due to manufacturing tolerances, roughness in vertically plated walls, and the gain tolerance of the reference antenna used for calibration. Notice that the simulations very accurately predict the 3 dB beamwidth varying between 45° at the low-end and 30° at the high-end, as well as the locations of pattern nulls and maxima. Also, the measured gain and 3 dB beamwidth of the two beams produced are similar, indicating that they are highly symmetrical as expected.
Also note that because an air-filled rectangular coaxial transmission line is used, ohmic losses are small and thus radiation efficiency is high.

![Figure 6.23: Measured and simulated realized gain and 3 dB beamwidth of the fully integrated two element subsystem and both beams.](image)

To fully evaluate the proposed prototyping approach, the effects of PCB manufacturing tolerances are studied on the directly integrated two element subsystem. The manufacturing tolerances expected to be the biggest issue are the widths of central conductor and air gap of each section of rectangular coaxial line, as well as board thickness. To test these effects, the width of each central conductor line section is varied by ±5% while keeping the total width of the coaxial cross section constant. This is done for each line shown in Table 6.1, and also for the 50 Ω feed lines. The effect of varying board thickness by ±10% [161] is also studied. The impact of these tolerances on the complete subsystem’s reflection coefficient, port to port coupling, gain, null depth, 3 dB beamwidth and location of the E plane beam maximum as compared to their nominal values are shown in Figure 6.24. As seen, the reflection coefficient of the antenna array and coupling between antenna ports vary from +2.5 dB to -10 dB. Far-field parameters including gain, null depth, 3 dB
beamwidth, and E-plane beam maxima vary by $\pm 2$ dB, $+4$ dB to $-10$ dB, $\pm 5^\circ$, and $\pm 6^\circ$, respectively. Note that while the reflection coefficient and antenna port coupling variations are larger in absolute sense, they are still below $-10$ dB over the entire 4 to 8 GHz bandwidth. Similar observation holds for the maximum gain, which remains above 9 dBi over the bandwidth for most cases, and null depth, which remains below $-10$ dBi. An important impact of tolerance to assess is the location of the beam maxima in the E-plane. The relatively small variation in the location of this maximum and 3 dB beamwidth shows that manufacturing tolerances will have minimal effect on the main lobe of the radiation pattern. While the maximum gain of the subsystem varies more than desired, the stability of the main lobe location and beamwidth confirms the suitability of the stacked PCB approach for prototyping 3D surface micromachined components and systems. In the frequency range of this particular prototyping subsystem, microstrip is the prevalent transmission line for many components. However, using transmission lines such as microstrip or stripline for prototyping coaxial-based 3D micromachined systems will not provide much useful information on their electrical performance and/or mechanical stability. In addition, it is quite possible that stand-alone components and integrated subsystems built using the stacked PCB approach, including the herein presented 90° hybrid and two element antenna array, may be good candidates for some applications in their own right. Considering inherent properties of an air filled rectangular coaxial line such as pure TEM operation and low loss, the proposed PCB stacking approach can be used for wideband direction finding or communication applications within UHF / Ka-band range.
Figure 6.24: Manufacturing tolerance effects on the performance of the integrated two element antenna array subsystem obtained with HFSS. Shown are (a) the reflection coefficient and coupling between input ports, (b) maximum and null gain, and (c) 3 dB bandwidth and location of the E plane beam maximum.
6.3 Prototyping using 3D Printing

While PCB prototyping of RF and millimeter wave subsystems can be used for many components, there is still a lead time in terms of sending and receiving the fabricated board. Many PCB fabricators offer special 24 hour turnaround times for time-sensitive prototypes, but these fast turnaround conditions are also more costly than standard turnaround times of one to two weeks. In addition, each PCB foundry has different capabilities, and some have minimum order sizes. To counteract some of these issues, and also to construct prototypes of RF and millimeter wave systems incapable of fabrication using PCB, 3D printing may be used [162]. As described in Chapter 1, 3D printed objects can be constructed using a variety of processes and materials. If the material used is not electrically conductive, electroplating can be used to make the deposit metal at critical locations for intended RF use. An advantage of this process is the decrease in cost of desktop 3D printers, which allows for the designer to rapidly prototype antennas and complete systems in house.

The small aperture horn is printed using a commercially available extrusion deposition printer from MakerBot [163], which uses acrylonitrile butadiene styrene plastic for fused deposition modeling. The antenna is 1” height, with a 1.85” diameter. The aperture has a width of 0.35”, ridge width of 0.072” and a ridge gap of 0.285”. The quad ridge aperture is transitioned linearly to a WRD-180 input. In addition to being printed, the antenna is also constructed using wire EDM from Custom Microwave [164]. The computer model of the 3D printed antenna, and actual 3D printed antenna and EDM
antenna are shown in Figure 6.25. Note that the 3D printed antenna has nearly 90% less mass from the EDM antenna, weighing in at 0.05 kg vs. 0.46 kg.

Figure 6.25: Computer model and fabricated small aperture horns. Shown are (a) the computer model, (b) un-plated 3D printed horn, (c) plated 3D horn and (d) EDM horn.

To facilitate measurement, the 3D printed model has two unique features. First, to limit the amount of mechanical post-processing needed, threaded holes on the opposite side of the radiating aperture (which are needed for the double ridge waveguide to coaxial adapter) are not used. Rather, the screws come through the body of the antenna to take
advantage of threaded holes present on the adapter itself as seen in Figure 6.25(a). If these holes are left open, they can impact surface currents present on the antenna, and negatively affect performance. Therefore, the holes are covered with copper tape, as seen in Figure 6.25(c). Second, the flange needed for far field measurements is directly attached to the horn. This is unlike the EDM horn, where a separate flange (along with connecting holes and alignment pins) are needed for far field measurement.

The impedance is measured up to 40 GHz on an Agilent E8363B PNA using a standard 2.4 mm SOLT calibration, and simulated and measured results for the EDM and 3D printed horn are shown in Figure 6.26. As seen, VSWR for the 3D printed horn is below 2 over the measured 19 to 40 GHz range. Note the excellent agreement with the EDM horn, and good agreement with simulated values over this frequency range.

![Figure 6.26: Measured and simulated impedance of the small aperture horn.](image)

Far field measurements are undertaken for both the EDM horn and 3D printed horn with either a WR-42 or WR-28 open ended waveguide horn as a probe. Radiation patterns for both antennas, compared with simulated values, are shown in Figure 6.27. As seen,
radiation patterns for the 3D printed horn compare favorably to simulated and EDM horn patterns. Due to small aperture size at these frequencies and symmetrical nature, the beamwidth is wide in both E and H planes. Increased rippling is noticed for the measured patterns, and this is attributed to the low gain of the probe and measured horns in the far field range.

Measured and simulated boresight gain and E and H plane 3 dB beamwidths for the corrugated cone stub are shown in Figure 6.28 and 6.29, respectively. As seen, measured boresight gain is between 7 and 12 dBi. The difference between simulated and measured gain is less than 1.5 dB throughout the band. Measured E plane 3 dB beamwidth is between 110° and 40°, while the H plane is between 75° and 50°. Note the large dip in E plane 3 dB beamwidth at 20 GHz due to the chosen aperture dimensions. Overall, the good correlation between the 3D printed horn and wire EDM horn confirms that 3D printed is able to be used for prototyping future millimeter wave antennas and components.
Figure 6.27: Measured and simulated E and H plane radiation of the small aperture horn.

Figure 6.28: Measured and simulated boresight gain the small aperture horn.
6.4 Summary

Low-cost prototyping of RF and millimeter wave front-end systems using PCB prototyping and 3D printing is introduced. First, a stacked PCB fabrication approach is proposed for fast prototyping of rectangular coaxial components and subsystems aimed for fabrication using 3D micromachining processes. A 4 to 8 GHz two element tapered slot antenna array with directly integrated 90° hybrid and its standalone components are designed to demonstrate this approach. The individual 90° hybrid has measured amplitude misbalance of less than ± 2 dB, phase misbalance of less than ± 7°, and reflection coefficient and port to port coupling below -10 dB over the full octave bandwidth. The two element tapered slot antenna array has reflection coefficient < -10 dB over majority of 4 to 18 GHz band with measured far-field as predicted by the theory. The directly integrated subsystem containing these two components shows properly formed radiation patterns, realized gain from 9 to 11 dBi, and 3 dB beamwidth from 30° to 45° for both simulation and measurement. Excellent correlation between the PCB prototype and a Ka band
micromachined subsystem is also observed. Obtained results indicate that the proposed approach can save time and cost in the development of more complex microwave and millimeter wave systems based on micro-coaxial and other 3D lines. Moreover, with some design and assembly optimization, this process may be able to produce good quality, very low-cost components and integrated subsystems on their own.

Secondly, 3D printing is used to prototype an 18 – 45 GHz small aperture horn, and its performance is compared to a conventionally fabricated wire EDM horn. As shown, the 3D printed small aperture horn has VSWR < 2, gain from 6 dBi to 12 dBi, and wide beamwidth radiation patterns over an 18 GHz to 45 GHz range. The results compare favorably to simulated values and the referenced EDM horn.
CHAPTER 7
Summary, Contributions and Future Work

7.1 Summary

The analysis, design, fabrication and measurements of novel wideband passive front ends for millimeter wave electronic warfare systems have been researched in this thesis. For electronic support applications, surface micromachined direction finding with multiple element antenna arrays are demonstrated. A wideband tapered slot antenna is designed and measured first, and is used as an array element for the subsequent arrays. VSWR < 2 and boresight gain > 4 dBi are seen over a 43 – 140 GHz range. Amplitude only and amplitude/phase two element direction finding front ends operating over 18 – 36 GHz and 75 – 110 GHz have also been demonstrated. It is shown that these two element front ends may be used in a stand-alone system or as part of a more complex aperture, and may provide good accuracy and range. The designs of arrays with more than two elements with integrated beamformers are also considered for direction finding over W band and beyond. Both Butler matrix and Rotman lens topologies have been tested with VSWR < 2 and gain > 8 dBi with expected radiation patterns. It is seen that the Butler matrix array has radiation patterns that change over frequency due to increasing directivity of the array, while the Rotman lens array has inherently consistent beam location due to its true time delay beamforming. In addition, the Rotman lens has larger physical size and insertion loss than the Butler matrix fed array.
A dual polarized high power capable front end for electronic attack over an 18 – 45 GHz band utilizing a custom double ridge waveguide cross section is proposed. To combine two polarizations into the same radiating aperture, an orthomode transducer (OMT) using the same double ridge waveguide cross section is engineered. To provide greater flexibility in performance, several matching stubs used in the turnstile junction section of the OMT are designed, fabricated and tested. Each matching stub has dual polarized inputs with return loss > 9 dB and isolation > 35 dB over the band. A modular horn section is then proposed to increase system diversity and capabilities. The modular horns are directly attached to the small radiating aperture of the OMT and can be designed for performance criteria such as constant gain, beamwidth, etc. A 13 dBi, 15 dBi and 18 dBi constant boresight gain horns are developed, with measurements comparing well to simulations for all horns. Beamforming networks utilizing the proposed double ridge cross section are also developed. Specifically, an eight section branch guide coupler with measured return loss and isolation > 12 dB, amplitude misbalance < 5 dB and phase difference 90° ± 5° is demonstrated. A high power capable Rotman lens with simulated insertion loss of < 2 dB is also designed and measured. Measurements confirm consistent beam location and gain between 9 and 18 dBi.

Finally, two low cost prototyping methods for the herein designed millimeter wave electronic warfare front ends are investigated. A 4 – 18 GHz two element array fed by commercially available beamformers and 4 – 8 GHz two element array with integrated beamformer based upon the stacking of PCB boards are shown. Similar performance as compared to the scaled micromachined models is demonstrated. A 3D printed small aperture horn is compared with a horn fabricated via wire EDM, and measured results show similar performance with a ten-fold reduction in cost and weight. This thesis has
shown that wideband millimeter wave front ends for electronic warfare can now be realized and used in today’s ever changing threat environment.

7.2 Original Contributions

- A 43 – 140 GHz micromachined tapered slot antenna was designed and fabricated. A thorough parametric study is undertaken to better understand this configuration and to ensure good performance in the available fabrication process. The antenna design is such that it can be easily adopted in various array based electronic warfare application.

- Two element micromachined amplitude only and amplitude/phase front ends for monopulse direction finding are designed, fabricated and measured over an 18 – 36 GHz and 75 – 110 GHz bands. All designs have VSWR < 2, with a field of view > 40° for amplitude only configurations and > 60° for amplitude/phase configurations.

- The first surface micromachined monolithic Rotman lens is designed for operation from 65 to 170 GHz. To the author’s best knowledge, this is the widest non-optical Rotman lens available in literature. Moreover, the designed lens monolithically integrates 5 different types of transmission lines, and achieves low loss and excellent pattern stability.

- A flexible orthomode transducer for 18 – 45 GHz dual polarized electronic attack applications is reported. Five different matching stub sections are fabricated and measured, with results comparing favorably to simulations. These matching stubs can be readily interchanged and used based on desired performance characteristics.
A modular horn extension for the 18 – 45 GHz orthomode transducer is proposed to achieve flexible options for ever changing operational requirements. Based upon thorough parameter studies of the horn extension, a 13, 15 and 18 dBi constant boresight gain apertures are developed.

A double ridge waveguide fed Rotman lens operating from 18 – 45 GHz is designed. By carefully selecting lens parameters, insertion losses comparable to rectangular waveguide based configurations can be achieved over a much wider bandwidth.

A novel prototyping method for micromachined electronic warfare systems is introduced. PCB boards are used to mimic the stacking process of typical micromachining processes. A 4 – 18 GHz two element antenna array and 4 – 8 GHz two element amplitude only subsystems are demonstrated using this method.

Final, the use of fused deposition modeling, or 3D printing, for prototyping millimeter wave apertures is reported. It is shown that a horn can have similar impedance and far field performance as compared to a wire EDM aperture with a tenfold reduction in cost and weight.

7.3 Future Work

The work presented here can lead to new and interesting research:

- The direction finding front ends in this thesis assumed perfect interconnects and no noise from back end circuitry, such that the calculated angle of arrival depended only on far field radiation patterns. In reality, this is not the case. The next step for these front ends is to integrate them with back end circuitry, with a focus on a fundamental understanding of interconnections between the front end and back end circuitry.
Future work in increasing the number of array elements for millimeter wave electronic warfare front ends will lead to higher directivity at the increase of a more complicated beamformer.

A fundamental understanding of high power operation for the electronic attack systems presented in this thesis is needed. This includes electromagnetic, thermal and mechanical simulations and experimental studies.

A fabrication method for double ridge waveguide orthomode transducers other than the split block construction used in this thesis is needed to prevent the seams and asymmetries that negatively affects isolation between the orthogonal ports.

Further investigation of how to use the stacking of PCB boards to fabricate not only electronic warfare front ends but subsystems for applications such as communications.

The further use of 3D printing for electronic warfare front ends. Studies can be undertaken to understand performance limitations associated with this fabrication technique.

The design of additional beamforming network components using the custom double ridge waveguide cross section in the 18 – 45 GHz band is needed. Specifically, phase shifters, polarizers, gratings, 180° hybrids, etc.
References


References:


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Custom Microwave, Custom Microwave, Inc., Longmont, Colorado, USA.
APPENDIX A

X through Ku band OMT With Integrated Gain Stabilized Horn

A.1 Introduction

Orthomode transducers (OMTs) are one of the key components in dual-polarization antenna feeds for applications such as communications [1], remote sensing [2] and radio astronomy [3]. OMTs are four-electrical port devices used to separate or combine two orthogonal polarizations at/to a common port. The bandwidth of these devices is dependent on the achievable structural symmetry which can suppress higher order modes [4]. Several topologies have been considered to achieve wider than usual bandwidth OMTs. These can be broadly classified into two primary groups. The first is based on the Boifot junction, which uses a septum in the branching region, capacitive compensation pins and both longitudinal and side wall ports [4] – [5]. The second group is based on the turnstile junction that uses four longitudinal ports and a concentric matching stub within a waveguide [6] – [7]. These designs have advantage of not needing a septum or pins to achieve wideband performance. However, even though the use of these design techniques is well understood, it is quite challenging to demonstrate an OMT with an octave or more bandwidth. This is because the performance is degraded by higher order mode excitation in the branching structures, which are needed for the orthogonal polarizations.

The OMT introduced in this section utilizes a double ridge bifurcation and single ridge waveguide turnstile junction to overcome the above discussed limitation and achieve good performance over a 7.5 to 20 GHz frequency range. Compared with previously published OMTs designed to operate within this frequency range, the herein discussed
device has similar isolation (> 36 dB) but it operates over wider instantaneous bandwidth, as shown in Figure A.1. The return loss is somewhat higher; however, it is still acceptable for many applications (>10 dB throughout the bandwidth). The OMT is built from constitutive parts machined using standard low-cost CNC machining process. Comparison between simulated and measured results confirms the good performance of the introduced OMT. Straightforward integration with a quad ridge horn antenna is also shown to be feasible, specifically with a 15 dB gain stabilized horn shown herein.

![Bandwidth comparison](image)

**Figure A.1:** Bandwidth comparison between this work and several comparable OMTs. Shown is the reference and simulated (measured) isolation in dB.

This Appendix is organized as follows:

- Section A.2 discusses the design of a 7.5 GHz to 20 GHz orthomode transducer, and concentrates on the development of key features in the turnstile junction fed quad ridge output.
- Section A.3 presents the attachment of a previously designed 15 dB gain stabilized horn to the quad ridge output of the orthomode transducer. Measurements of both the standalone gain stabilized horn and the complete system confirm good far field performance of the system.
A.2 X thru Ku band Orthomode Transducer

The full model of the designed OMT is shown in Figure A.2. As seen, the OMT is composed of three basic components/parts: a double ridge waveguide bifurcation, a single ridge waveguide turnstile junction and necessary E plane bends in both single and double ridge waveguide lines. It is noted that ports P1 and P2 are extended out to provide the necessary spacing to accommodate standard WRD-750 flanges, but are not otherwise part of the functional design of the OMT. All simulations, analysis and design of individual parts as well the complete device are conducted using HFSS [9].

![Figure A.2: Full model of the OMT based on a double ridge waveguide bifurcation and single ridge waveguide turnstile junction.](image)

To obtain a better match throughout the operating bandwidth, a modified double ridge waveguide cross-section based on scaled WRD-180 is developed. The final designed cross section has a width of 16.16 mm, height of 7.27 mm, with the ridge gap and ridge width being 2.5 mm and 3.88 mm, respectively. In comparison, the commercially available cross section WRD-750 for this bandwidth has a width, height, ridge gap and ridge width of 17.55 mm, 8.15 mm, 3.45 mm and 4.39 mm respectively. Note that the reduced ridge gap will contribute to slightly increased losses in the system. However, the narrower ridge gap
leads to increased bandwidth in the system, with higher order modes not propagating until 20.5 GHz. In addition, because the phase constant between the first order modes in both waveguides is quite similar, WRD-750 waveguide can be connected to this custom cross-section with a return loss better than 17 dB.

The double ridge waveguide bifurcation is inspired by waveguide based designs from [10] and [11]. Therein, the input rectangular waveguide had gradual steps to provide a better impedance match to the output ports. However, the use of any impedance transformations can be avoided in this design by splitting the double ridge waveguide input into two single ridge waveguides. This results in a more compact design, with a final length of 4.67 mm, width of 16.16 mm and height of 9.7 mm. Note there is a gap of 2.44 mm between the single ridge waveguide outputs. The final design and obtained match are shown in Figure A.3. As seen, the reflection coefficient is better than -30 dB over the band, with insertion loss of less than 0.006 dB for aluminum platform.

Figure A.3: Structure and simulated reflection coefficient of the double ridge waveguide bifurcation.
The design of a single ridge waveguide turnstile junction is based on a similar approach for a rectangular waveguide device from [12]. A square quad ridge waveguide is used at the output of the junction for ease of fabrication. The square quad ridge waveguide has a width of 13.44 mm, ridge width of 3.9 mm and ridge gap of 4.66 mm. A pyramidal stub is located at the base of the turnstile junction to enhance its broadband performance. The designed pyramid has a width of 4.8 mm and height of 2.4 mm. Final height, width and length of the turnstile junction are 30.5 mm, 30.5 mm and 11.43 mm, respectively. The simulated match is shown in Figure A.4. As seen, the reflection coefficient is below -10 dB over the entire bandwidth. While better mismatch performance is typically desired in an OMT, note that the reflection coefficient is below -15 dB from 9.15 GHz to 18.2 GHz, leading to very good performance over nearly an octave of bandwidth.

![Figure A.4: Structure and simulated reflection coefficient of the single ridge waveguide turnstile junction.](image)

Having designed the turnstile junction and bifurcation, the only remaining task in assembling the full OMT is to integrate the bifurcation structure with the orthogonal port. A positioning conflict arises if the bifurcation structure is simply copied and rotated by 90°,
so lateral repositioning of both bifurcations is necessary, which introduces feed asymmetry. This asymmetry can be compensated assuming that the E-plane bends occurring between the bifurcation and turnstile have very good match. Since these bends have reflection coefficient below -30 dB across the entire band, their effect was found to be negligible. The compensation then consists solely of routing additional path lengths in the turnstile feeding waveguides to equalize their length.

To complete the design, the double ridge feed lines are transitioned out to create enough space to attach dual WRD-750 coaxial adapters from Space Machine [13]. The final size of the device is 108 mm × 108 mm × 95 mm, the majority of which is fixturing overhead. The OMT is constructed using the split-block CNC machining, and held together using screws as seen in Figure A.5. The quad ridge output is terminated using an available quad ridge horn with return loss greater than 15 dB over the entire bandwidth.

An Agilent N5246A VNA is used for S-parameter measurements, and obtained reflection coefficient and isolation are compared to simulated values in Figures A.6 and A.7, respectively. As seen, the measured reflection coefficient is less than -10 dB over the 7.5 to
19 GHz range. Above 19 GHz, the performance of the WRD-750 coaxial adapter noticeably deteriorates, leading to the observed discrepancy. Measured isolation is better than 36 dB over the frequency band, with simulated isolation being better than 55 dB. The discrepancy in isolation is due to the small air gaps between the aluminum blocks. The effect of these gaps can be minimized through the use of a gasketing material or mechanical lapping of the mating surfaces. While doable, these would increase the prototyping cost and was not actively pursued at this time.

Figure A.6: Measured and simulated reflection coefficients of the OMT at both ports.

Figure A.7: Measured and simulated isolation of the OMT.
A.3 15 dB gain stabilized horn attachment for OMT

Note that the developed OMT needs to be connected to a quad ridge horn antenna or other compatible aperture for its practical use. The type of quad ridge horn used can be easily interchanged depending on the application. For example, the horn can be designed for high directivity, stable gain or beamwidth, or small aperture size. For this specific application, a gain stable antenna is developed [14] for direct integration with the quad ridge output of the OMT. Similar to the design in Chapter 4, a small aperture section is first developed, and then a horn extension is added onto the end of this. However, for this specific horn extension, a filleted radius blend between the small aperture and extension sections is used. While this prevents the horn from being modular as in Chapter 4, it does provide for a smoother mode transition between the two sections. Parameters studied for this configuration include the total length and flare angle of the additional horn section, and also the ridge and waveguide wall blend radius. The final size of the horn is 108 mm × 108 mm × 129 mm, with a 50 mm blend radius between the two sections. The final model of the gain stabilized horn is shown in Figure A.8, and measured and simulated VSWR and gain are shown in Figures A.9 and A.10, respectively.
As seen, measured VSWR is less than 1.6 over a 7.5 GHz to 19 GHz band. Above 19 GHz, reflection and losses from the WRD-750 coaxial adapter increase, limiting the effective measurement band. Measured gain of the horn is 15 dBi ± 0.75 dB over the same bandwidth. Both measurements and simulations have similar trends over the frequency band. The gain stabilized horn is connected to the quad ridge output of the OMT and
screwed down. The final model is shown in Figure A.12, and E and H plane radiation patterns in Figure A.13. Measured E and H plane radiation patterns compare well to simulations. Note that E plane rippling is reduced compared to the gain stabilized horn models designed in Chapter 4. This is because of the fillet blend between the small aperture and horn extension sections, providing a smoother transition between the two.

Figure A.10: Measured and simulated gain of the gain stabilized horn.

Figure A.11: Fabricated OMT with attached gain stabilized horn.
To further show the good performance of the system, measured and simulated E plane 3 and 10 dB beamwidths, H plane 3 and 10 dB beamwidths and gain are given in Figures A.13, A.14 and A.15, respectively. E plane 3 dB beamwidth varies between 20° and 40° over the band, while H plane 3 dB beamwidth between 25° and 35°. Note the increase in E plane 3 dB beamwidth at approximately 12 GHz due to the presence of higher order modes in the horn aperture. Measured directivity compares well to the simulated gain as
seen in Figure A.15, and is 15.5 dBi ± 0.5 dB. However, the measured gain does not match up as well, having a gain of 14.5 dBi ± 1.5 dB. This is due to increased losses coming from seams and rough surfaces in the OMT that are not present in the simulated model. As discussed earlier, these losses can be minimized through the use of a gasketing material or mechanical lapping of the mating surfaces.

Figure A.13: Measured and simulated 3 and 10 dB E plane 3 dB beamwidth of the OMT with gain stabilized horn. Note the measurements of both polarizations of the OMT.

Figure A.14: Measured and simulated 3 and 10 dB H plane 3 dB beamwidth of the OMT with gain stabilized horn. Note the measurements of both polarizations of the OMT.
A.4 Summary

The performance of a wideband OMT based on a double ridge waveguide bifurcation and single ridge waveguide turnstile junction is presented. The OMT has measured return loss greater than 10 dB and isolation greater than 35 dB over a 7.5 to 20 GHz bandwidth. A 15 dB gain stabilized horn is then designed as an attachment to the quad ridge output of the OMT, and has VSWR less than 1.6 over the same bandwidth. When combined, measurements confirm good far field patterns with E and H plane 3 dB beamwidths of 20° to 40° and 25° to 35° over the band, respectively. Measured directivity is 15.5 dBi ± 0.5 dB and measured gain is 14.5 dBi ± 1.5 dB. This is due to increased losses in the fabricated OMT, and future work will concentrate on reducing these losses.
References


