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**Polymer-based Additive Manufacturing for Microwave
and Millimeter-wave Applications**

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Declaration of Originality

I herewith certify that to the best of my knowledge; the content of this dissertation is my own work. All the material which is not my own work has been properly acknowledged. The research presented in this this dissertation is conducted by me under the supervision of Prof. Stepan Lucyszyn.

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Abstract

The objective of this thesis is to investigate the effectiveness of using various low-cost 3-D printing technologies for microwave and mm-wave applications. Different types of 3-D printers are used, and the corresponding printing and post-processing procedures are developed to design and fabricate microwave, mm-wave and quasi-optical components and subsystems. In addition, a bespoke mm-wave single-pixel camera is built for benchmarking a commercial camera, which will be used for characterizing 3-D printed quasi-optical components.

Chapter 1 provides an overview of current trends in 3-D printing for microwave and mm-wave applications. A literature survey of components relevant to this dissertation is included.

Chapter 2 is an overview of current polymer-based 3-D printing techniques. General information of common additive manufacturing technologies such as fused deposition modelling, vat photopolymerization and material jetting is shown in this chapter.

Chapter 3 reviews details on setting-up FDM and vat photopolymerization 3-D printers. Important printing parameters such as layer thickness and build-orientation are discussed in this chapter. Also, detailed post-processing procedures used for fabricating devices described throughout this thesis are given. Finally, a methodology for setting up a safe lab-environment for 3-D printing and post-processing is presented.

Chapter 4 introduces a fully 3-D printed 4-elements phased-array operating at Ku-band (12 to 18 GHz). A combination of commercial-level material jetting and a low-cost desktop FDM printer is used to fabricate the antenna body parts and dielectric phase shifters, respectively. Measured performances match well with theoretical and simulated results showing the effectiveness of using the 3-D printing for this application.

Chapter 5 covers the design, fabrication and measurement process of quasi-optical components such as horn antennas, mirrors, and RAM for G-band (140 to 220 GHz) applications. In addition, A simple quasi-optical subsystem is 3-D printed to be used for integrating and testing the quasi-optical components.

Chapter 6 follows up on the work from Chapter 5 and demonstrates a multi-channel subsystem that integrates various quasi-optical components. Newly customized components such as beamsplitters and convex lenses are integrated with the subsystem.

Chapter 7 presents a commercial mm-wave camera benchmarking using a custom-built raster-scanning camera. The mm-wave camera is benchmarked as it is one of the most effective tools when characterizing 3-D printed quasi-optical components such as lens and mirror. The custom-built mm-wave camera will be used for future studies on characterizing or finding defects on various 3-D printed quasi-optical components.

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List of Symbols

A_{max}	maximum internal aperture area	mm^2
$\mathcal{A}$	absorptance	
$\mathcal{A}_{p-s}$	embedded polarizer power absorptance	
a	waveguide internal cross-sectional width	mm
b	waveguide internal cross-sectional height	mm
c	speed of light	m/s^2
D	mirror Diameter	mm
D_e	theoretical directivity single antenna element	
D_i	theoretical directivity of isotropic element linear array	
D_A	boresight directivity of complete phased-array antenna	
D_{STD}	standard gain horn maximum aperture size	
D_{PAA}	phased-array antennas maximum aperture size	
D_{HA}	largest internal aperture dimension	
d	distance between antenna elements for phased-array	mm
d_{px}	pixel width	mm
d_{py}	pixel height	mm
δ	skin depth	μm
E_p	extinction ratio	
ε_r	dielectric constant	
ε_{reff}	effective relative permittivity	
ε_t	Aperture taper efficiency	
ε_{ph}^H	Phase efficiency	

f	frequency	GHz
f_c	TE upper cut-off frequency	GHz
f_{min}	minimum operating frequency	GHz
f_{pa}	parent Focal Length	mm
f_{eff}	reflected effective focal length	mm
f_{pl}	focal path length	mm
G	boresight antenna gain	
G_R	reference boresight antenna gain	
G_T	boresight gain of antenna under tests	
g	dielectric insert penetration Depth,	mm
η_o	free space impedance	Ω
θ	phased-array teering angle	$^\circ$
θ_p	polarizer rotation angle	$^\circ$
h	polarizer strip separation distance	μm
I	beam intensity	W/mm^2
I_0	peak beam intensity	W/mm^2
λ	wavelength	mm
λ_0	wavelength in free space	mm
λ_g	wavelength in waveguide	mm
λ_{min}	minimum wavelength within operating frequency band	mm
μ_o	free space permeability	N/m
n	refractive index	
ξ	the fractional aperture distribution phase error	
P	polarization efficiency	

p	beamsplitter's lattice constant	
ρ_o	bulk resistivity	$\Omega \cdot \text{m}$
R_s	line-of-sight aperture-to-aperture separation distance	mm
R_{ff}	radiating far-field (Fraunhofer) distance	mm
$\mathcal{R}$	reflectance	
$\mathcal{R}_{p-s}$	embedded polarizer power reflectance	
r	lens radius	mm
S_{11}	forward reflection coefficient	
S_{12}	reverse transmission coefficient	
S_{21}	forward transmission coefficient	
S_{22}	reverse transmission coefficient	
σ_o	bulk conductivity	S/m
T_m	metallization thickness	μm
$\mathcal{T}$	power transmittance	
$\mathcal{T}_n$	normalized transmittance of the beamsplitter	
T_{p-s}	embedded polarizer power transmittance	
T_p	de-embedded polarizer power transmittance	
$\tan\delta$	dielectric loss tangent	
ϕ	phase angle	o

List of Abbreviations

ABS	acrylonitrile butadiene styrene
BPF	band-pass filter
CAD	computer-aided design
CE	cyanate ester
CMOS	complementary metal-oxide-semiconductor
CNC	computer numerical control
DUT	device under test
EM	electromagnetic
FDM	fused deposition modelling
FPA	focal plane array
FWHM	full width at half maximum
HEPA	high-efficiency particulate absorbing
HMMH	horn-mirror-mirror-horn
IPA	isopropanol
IR	infrared
LCD	liquid crystal display
LED	light emitting diode
MPRWG	metal-pipe rectangular waveguide
MSLA	masked stereolithographic apparatus
NEP	noise equivalent power
OAPM	off-axis parabolic mirror
PEC	perfect electrical conductor
PLA	polylactic acid
QS	quasi-optical

RAM	radiation absorbent material
RF	radio frequency
OPL	optical path length

List of Frequency Band and Waveguide Designation

X-band	WR-90	-	8 to 12 GHz
Ku-band	WR-62	-	12 to 18 GHz
Ka-band	WR-28	-	26.5 to 40 GHz
Q-band	WR-22	-	33 to 50 GHz
V-band	WR-15	-	50 to 75 GHz
E-band	WR-12	-	60 to 90 GHz
W-band	WR-10	WM-2540	75 to 110 GHz
G-band	WR-5	WM-1295	140 to 220 GHz
-	WR-3.4	WM-864	220 to 330 GHz
-	WR-2.2	WM-570	325 to 500 GHz
-	WR-1.5	WM-380	500 to 750 GHz
-	-	WM-250	750 GHz to 1.1 THz

1 Trends on Polymer 3-D Printing for Microwave & Millimeter-wave applications

The contents in this literature review are based on the published article by Shin *et al.* [1-3]. Additive manufacturing using polymer-based 3-D printing is an emerging technology that is finding its way from academic research to commercial exploitation. Over the past few years, companies have been adopting additive manufacturing techniques as a standard means of rapid prototyping and small-batch production. For example, 3-D printing has inherent advantages with fast turn-around, weight reduction (employing unique topological solutions) and waste reduction, when compared to subtractive manufacturing. These significant advantages already play a key role within the industries such as aerospace and high-performance automotive. This enabling technology has already proven itself with metal-pipe rectangular waveguides (MPRWGs) and associated components.

Polymer-based 3-D printing is now showing great potential for the rapid prototyping of passive components and circuits for microwave-to-terahertz (THz) applications. For example, Timbie *et al.* demonstrated a 3-D printed metal-pipe rectangular waveguide (MPRWG) and corrugated horn antenna at W-band (75 to 110 GHz) [4]. Three years later, Bieren *et al.* reported a MPRWG and diagonal horn antenna at WR-3.4 band (220 to 330 GHz) [5]. The following year, detailed investigation for implementing high performance 3-D printed MPRWGs were reported; covering both X-band (8 to 12 GHz) and W-band, while also demonstrating a high performance E-plane split-block iris-coupled 6th order band-pass filter (BPF) at W-band [6]. Since then, the same team have moved 3-D printed MPRWGs up in frequency; covering the WM-380 band (500 to 750 GHz) and WM-250 band (750 GHz to 1.1 THz) [7].

These early works [4]-[7] are non-reconfigurable, in the sense that the reported components are neither mechanically nor electrically tunable. In order to expand the functionality for 3-D printing, in addition to metal-plated structures, important polymer-based building materials has been accurately characterized. For example, a relatively low loss acrylonitrile butadiene styrene

(ABS) material was characterized over X-band [8] and Ku-band (12 to 18 GHz) [9]. This build material was then employed in mechanically-tunable 3-D printed dielectric-insert phase shifters at both X-band [10] and Ku-band [1]. In addition, carbon-loaded conductive polylactic acid (PLA) was also characterized across both X- and Ku-bands [9]. This build material was then employed in a mechanically-tunable rotary vane attenuator at Ku-band [9].

While mechanically-tunable 3-D printed components are interesting, there are still reliability issues. A passive-active hybrid technology was demonstrated, having 3-D printed waveguide flanges, delay lines and power splitters/combiners and, more significantly, with integrated high-resistivity silicon implants and laser diodes; creating an electronically-tunable I-Q vector modulator at WR-2.2 band (325 to 500 GHz) [11].

Integrating multiple 3-D printed components together offers the important advantage of removing impedance mismatch losses between flanges. For example, both Rohrdantz *et al.* [12] and later Dimitriadis *et al.* [13], demonstrated a non-tunable 4-element antenna array at Ka-band (26.5 to 40 GHz). In contrast, a mechanically-tunable 3-D printed beam-steerable 4-element phased-array antenna at Ku-band has recently been demonstrated [1].

In addition to the advances being demonstrated with integrated waveguide-based subsystems, Harris Corp. (USA) recently reported two commercial (conventionally-machined) open-ended waveguides that are coupled via a pair of 3-D printed lens, using an in-line alignment frame at lower millimeter-wave frequencies (*ca.* 30 to 100 GHz) [14], [15].

The following represents a review of 3-D printed components and subsystems of the types investigated in this thesis. It should be noted that for each, less than a handful of polymer-based 3-D printed examples could be found in the open literature for upper millimeter-wave (*ca.* 100 to 300 GHz) and sub-millimeter-wave/terahertz (0.3 to 3 THz) frequency applications.

1.1 Phased-Array Antennas

Phased-array antennas are widely used for many applications, such as in modern radar and space communications systems. Phased-array antennas have many identical stationary antenna elements that form a steerable beam. It requires multiple sets of highly customized components, such as transmission lines or guided-wave structures, power splitters, phase shifters, feed networks and radiating elements [16]. Because these components are usually produced in small batches, perhaps requiring expensive bespoke manufacturing tools and lengthy turn-around times, the high fabrication cost is a major disadvantage when conventional subtractive manufacturing techniques are used (e.g., milling and machining). In this subsection literature review of 3-D printed components that could be used for a phased-array antenna subsystem will be presented.

With conventional MPRWGs, components are manufactured as separate parts and assembled to make a complete subsystem, requiring many waveguide flange connections; each contacting pair introducing a degradation in performance. In contrast, 3-D printing technologies allow more complex 3-D structures to be manufactured. Also, the number of parts can be significantly reduced with 3-D printing, minimizing the number of flanges.

The potential of using 3-D printing and surface metallization for microwave, millimeter-wave and terahertz applications have been demonstrated by several proof-of-principle devices. X- and W-band rectangular waveguides have been successfully developed with fused deposition modeling (FDM) and (SLA) [7]. At Ku-band (12 to 18 GHz), different types of antennas, such as corrugated conical horn [17] and spline-profiled smooth horn antennas [18], were demonstrated using industrial FDM and SLA technologies, respectively. Using PolyJet 3-D printing, a G-band waveguide, standard gain horn antenna and parabolic mirror have been manufactured giving measured performances close to their commercial counterparts [19].

There is growing interest in utilizing the dielectric properties of polymer-based 3-D printed materials [8]. The dielectric properties of the polymers are easily tunable with composite filling materials and by varying infill densities.

For example, using FDM 3-D printers, an X-band (8 to 12 GHz) termination has been manufactured and tested using a carbon filled polymer [20], while a broadband microwave gradient-index lens with varying infill density has also been demonstrated [21].

Polymer-based 3-D printing has been used to demonstrate relatively complex devices with repeating features, such as waveguide slot arrays [22] [23]. A linear 10-element dielectric-filled millimeter-wave horn antenna array was fabricated using SLA, in 2008, and showed an operational bandwidth from 35.0 to 39.5 GHz [24]. A more complicated antenna array, which includes components such as a flange, E-plane dividers and horn antennas, has been developed using SLA [25]. Similarly, a 2×2 antenna array with complicated waveguide feed network and conical horn antennas was developed to operate at Ka-band (26.5 to 40 GHz) [26].

Several 3-D printed steerable antennas have been reported in open literature [27]-[28]. One was an S-band 6-element linear dipole array, made using multiple SLA-printed parts and needing six separate external feeds to provide steering [27]. Another was the electromagnetic simulation-only of a Ku-band frequency scanning slot array, having a scan angle of between -15.2° to $+15.4^\circ$ [28]. Finally, a resonant slot array has been demonstrated that employs mechanically moveable dielectric inserts to perform beam steering, employing SLA and FDM printing, respectively [29]. In that paper, the total beam steering angle is 0 to 8.5° , using a conventionally manufactures metal screw-spring mechanism, over a 1% fraction bandwidth at 21.5 GHz, having a worst-case VSWR of 2.9.

1.2 Mm-wave Horn Antennas

Horn antennas are commonly used in quasi-optical systems, as an interface between a waveguide medium and free space. As with conventional MPRWGs, horn antennas operating between 0.1 and 3 THz are very expensive. As a result, there is growing interest in low-cost manufacturing of horn antennas using 3-D printing technologies.

Metal-based 3-D printed antennas have demonstrated good performance. However, these relatively expensive manufacturing technologies still require post-processing to overcome inherent surface-roughness problems [30], [31].

With polymer-based 3-D printing, in 2011, Timbie *et al.* reported a split-block corrugated horn antenna operating over most of W-band using SLA technology [4]. Later, in 2016, Decrossas *et al.* demonstrated a W-band split-block corrugated horn antenna using a PolyJet 3-D printer [32]. In 2012, Macor *et al.* from SWISSto12 demonstrated a conical smooth wall horn antenna operating at 92.5 GHz using SLA technology [33]. Later, in 2014, Bieren *et al.* from the same team, reported a diagonal horn antenna operating in the WR-3.4 band (220 to 330 GHz) using SLA technology [5]. More recently, in 2018, Molaei *et al.*, demonstrated a PolyJet 3-D printed rectangular horn antenna operating at E-band (60 to 90 GHz) [34]. With all these examples, no direct performance comparison is made with an equivalent commercial counterpart. However, a comparison was made with an E-band SLA 3-D printed antenna (coated with silver spray paint) and compared with a commercial counterpart (electroformed with copper) and it was found that the gain was improved by an average of approximately 1 dB [35]. In summary, all these 3-D printing technologies have proven themselves suitable for manufacturing mm-wave and THz horn antennas using industrial-level 3-D printers.

In contrast, an ultra-low cost masked stereolithographic apparatus (MSLA) 3-D printer is employed here to fabricate ruggedized rectangular horn antennas, which are then deployed in fully 3-D printed subsystem demonstrators. Moreover, the performance of the modified replica antennas will be investigated and benchmarked against their commercial counterpart.

1.3 Parabolic Mirrors

Typically, a parabolic mirror is used to either tightly focus a collimated beam or to collimate a diverging beam, where beam divergence emanates from an effective focal point. With quasi-optical systems, 90° off-axis parabolic mirrors (OAPMs) are commonplace. These mirrors have very specific 3-D surface curvatures (cut across a circular paraboloid), requiring a minimum accuracy

dictated by a small fraction of the smallest wavelength of operation λ_{min} (e.g., $\lambda_{min}/20$). Therefore, for optical applications, accurate computer numerical control (CNC) machining and polishing processes are required. However, for (sub-)mm-wave applications, 3-D printing offers new opportunities, as demonstrated at 160 GHz using PolyJet technology [19]. At the same time, in Jul. 2018, Romeu *et al.* explored three different types of 3-D printing technologies (SLA, FDM and Multi Jet Fusion) for fabricating larger mirrors for operation at 100 GHz [36]. The measured beam profiles are comparable to their commercial counterpart. In 2019, Fullager *et al.* demonstrated a mirror similar to that previously demonstrated [19], but manufactured with a low-cost desktop SLA 3-D printer and tested at 530 GHz [37]. The measured beam profiles, determined using a THz camera (with sufficiently high source power and greater camera resolution), are comparable to their commercial counterpart.

In this work, the new Formlabs Low Force Stereolithography (LFS)TM 3-D printer was used technology [38], based on a type of SLA technology, to manufacture mass-reduced mirrors, which are then deployed in fully 3-D printed subsystem demonstrators and benchmarked against their commercial counterparts.

1.4 Radiation Absorbent Material

Radiation absorbent material (RAM) is used in anechoic chambers to reduce unwanted reflections, as it mitigates against adverse multi-path interference and standing-waves in line-of-sight paths. RAM can be manufactured with numerous combinations of polymers having magnetic and/or electric nanoparticle fillers. For example, Huber *et al.* [39] characterized 3-D printable polylactic acid (PLA) polymer composite materials in the microwave frequency range of 2 to 10 GHz: conductive PLA and magnetic PLA.

With lossy composite thermoplastic filaments, for FDM 3-D printers, there has been some attempts to fabricate free-space absorbers at microwave frequencies. For example, truncated pyramidal and frequency selective surface (FSS) metamaterial absorbers have been developed by Kjelgard *et al.*, operating from 2.6 to 4 GHz and 6 to 8.5 GHz, respectively [40]. Both absorbers are 3-D

printed using Proto-pasta conductive PLA [41], made by ProtoPlant Inc. (USA). In 2016, Arbaoui *et al.* characterized a carbon-loaded ABS polymer (TW-CON175BK, from Torwell Technologies) for use as a broadband microwave waveguide matched load termination across X-band [20].

In 2019, Petroff *et al.* developed a 3-D wedge-shaped RAM (broadband graded index absorber designed around a geometric approximation of the space-filling Hilbert curve) using carbon-loaded high impact polystyrene (HIPS) composite filament. The return loss (without metallic backing) is better than 22 dB, at normal incidence, in two different frequency bands: 63 to 115 GHz and 140 to 215 GHz [42].

In this work, Proto-pasta conductive PLA tessellating pyramidal-array RAM is FDM 3-D printed as zig-saw tiles and characterized from 140 to 220 GHz and benchmarked against their commercial equivalent. The bespoke RAM tiles line the inside of a quasi-optical housing, which are then deployed in fully 3-D printed subsystem demonstrators.

1.5 Grid Polarizer

In quasi-optical systems, a grid polarizer is a spatial filter that lets incident radiation of a specific polarization to pass through, while blocking radiation with other polarizations. Proto-pasta conductive PLA has also been used to develop an absorbing grid polarizer, for frequencies between 500 GHz and 2.7 THz. With strip widths and separation of 300 μm , extinction ratios better than 20 dB are demonstrated, with performance depending on the strip thicknesses (varying between 1 and 5 mm) [43]. However, this cannot be used as a beam splitter due to its absorbing nature. Ryu *et al.* has developed a grid polarizer with a type of SLA printer that uses the Texas Instruments digital micro-mirror projection system, with ultraviolet (UV)-curable acrylate-based resin [44]. This polarizer has a strip width and separation of 132 μm and 268 μm , respectively. The strips are sputter-coated with 100 nm thick gold, which lines the front face and 792 μm thick sidewalls. The maximum extinction ratio is 2.3 dB over their measured grid-effect frequency range from 2.94 to 2.50 THz [44]. The reason for this poor

performance is that they operate above the theoretical transverse-electric (TE) cut-off frequency of 0.56 THz.

1.6 Dielectric Lenses

As with the OAPM (in reflection mode), dielectric lenses (in transmission mode) are used to collimate and focus beams in (quasi-)optical systems. However, fabricating a custom lens using conventional CNC machining can be very expensive, especially if aspherical profiles and/or larger diameters are needed. Utilizing the inherent design flexibility offered by 3-D printing, a number of lens antennas and quasi-optical lenses have been demonstrated at mm-wave frequencies.

Using a PolyJet 3-D printer, Yi *et al.* reported dielectric lens antennas, having geometrically complex anti-reflection structures, operating from 50 to 70 GHz and 220 to 325 GHz [45]. Chudpooti *et al.* demonstrated hemispherical lens antennas operating from 220 to 320 GHz, manufactured using low-cost digital light processing (DLP) SLA 3-D printer technology [46]. Although these examples prove that these 3-D printing technologies are suitable for fabricating mm-wave dielectric lens antennas, they were not tested for quasi-optical beam collimation/focusing applications. Using a Multi-Jet-Modelling 3-D printing technology, three types of lenses (symmetric, elliptical-aspheric, planar-hyperbolic) were demonstrated for quasi-optical applications at 200 GHz; results show very accurate focal lengths [47]. More recently, in 2019, Friel *et al.* demonstrated a PLA-based FDM 3-D printed hyperbolic lens, with wavelength specific anti-reflective surface structures optimized for 60 GHz, having reasonably accurate focal points [48].

1.7 Quasi-optical Housing

In general, commercial machined breadboards and compatible mounts are used for (quasi-)optical prototyping; all having very high mechanical accuracy and requiring manual alignment, which is a very slow and expensive solution for the measurement of individual components and their integrated subsystems. Unlike with optical experiments, beam alignment in quasi-optical systems can be very

difficult to achieve, without laborious micrometer adjustments to the many individual positioners (each having multiple degrees of freedom). This is because visible laser alignment techniques are not always possible. Moreover, with multiple components interacting with the beam path, replacing any one component may require other components to be re-aligned. One solution is to create a special housing, with pre-aligned slots that can facilitate the ‘plug and play’ concept for individual components, without requiring re-alignment.

Currently, two examples of 3-D printed quasi-optical housing have been reported in the open literature, both published by the same group at Harris Corp. (USA) at 39 GHz in 2018 [14] and 60 GHz and 100 GHz in 2019 [15]. Their rigid plastic frame is 3-D printed, to create an air cavity for supporting two open-ended rectangular waveguides having a line-of-sight distance of fifteen wavelengths ($15\lambda_o$) in free space. In addition, two biconvex lenses and their supports are 3-D printed; the 100 GHz implementation also has 3-D printed rail alignment structures, designed for use with multiple interlock points on the frame. The material of choice for the lens, supports and alignment structures is Cyanate Ester 220 (CE 220), printed using a continuous liquid interface production (CLIP™) technique on the very expensive Carbon M1 3-D printer. This material is selected because of its flat dielectric properties between 20 to 57 GHz, measured using an open-resonator test system; the extracted band-average dielectric constant is 2.49 and loss tangent is 0.013 [14], [49]. Harris Corp. then state that CE 221 is used, as it is easier to print and post-process, while keeping the same electrical properties as CE 220. In their proof-of-concept demonstrators, they record the improvement in insertion loss, between the open-ended waveguides when lenses are introduced, which they term ‘focusing gain’. The reported values of measured insertion loss (without lenses) and focusing gain are given in Table 1, along with textbook calculations for Friis transmission path loss (without lenses): $-20 \log_{10}\{(a \times b)/\lambda_o R_s\}$, where $R_s \approx 15\lambda_o$ is the separation path length between the open-ended waveguides, both having internal width a and height b dimensions based on an assumption of the standard Electronic Industry Alliance (EIA) waveguide bands used. This ideal scenario also assumes 100% efficient radiation from both MPRWG apertures and so calculated path losses are

underestimates. It can be seen in Table 1 that the measured insertion loss (without lenses) is within 3 dB of prediction at 39 GHz. At this frequency, the quasi-optical lens system demonstrates a 14 dB port-to-port insertion loss for the coupling between two MPRWGs.

TABLE 1. Calculations and reported measurements [14] [15] of quasi-optical lens systems for coupling between open-ended waveguides.

Frequency (GHz)	Assumed Waveguide Band			Calculations		Reported Measured Results	
	EIA Designation	Internal Aperture Dimensions (mm)		Open-ended Waveguide Directivity (dBi)	Friis Transmission Path Loss Without Lenses (dB)	Insertion Loss Without Lenses (dB)	Focusing Gain (dB)
		a	b				
39 [14]	WR-22	5.69	2.84	5.37	35.1	32.5	18.6
60 [15]	WR-15	3.76	1.88	5.51	34.5	—	18.3
100 [15]	WR-10	2.54	1.27	6.54	32.4	—	22.1

1.8 Acknowledgements

The literature review presented in this chapter is based on journal papers published and submitted throughout this PhD [1],[2],[3]. Complete list can be found on Publications chapter at the end of this dissertation. Prof. S. Lucyszyn went over every word of the original article and provided valuable insights and feedback on the direction and presentation of the research work.

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2 Overview of Polymer-Based Additive Manufacturing Techniques

Additive manufacturing, also known as 3-D printing, is a fabrication technique that deposits single layers of materials to build three-dimensional structures. Compared to traditional manufacturing technologies such as machining and molding, 3-D printing offers several significant advantages that make it perfect for microwave and mm-wave applications. Conventional manufacturing techniques usually require specialized equipment such as machine tools and custom-made molds. These are ideal for mass production but can only be used for the same or similar components with specific dimensional and geometrical properties. Therefore, for prototyping or small-batch production of various component types, traditional manufacturing techniques are not very economical. Moreover, if the component's geometry is complex, the fabrication rises exponentially. In contrast, additive manufacturing technology allows users to design and fabricate geometrically complex structures without purchasing many different types of high-cost specialized machine tools. In most cases, a single 3-D printer can be used to manufacture various types of components, from rectangular waveguides to dielectric lenses. 3-D printer's affordability and ease of use allow fast turnaround in-house prototyping even in academic environments.

2.1 FDM

Fused deposition modelling (FDM) or fused filament fabrication (FFF) is one of the most common 3-D printing techniques. It was first introduced by Stratasys' co-founder S. Scott Crump in 1988, where he used a hot glue gun to melt and extrude thin layers of polymer to build 3-D structures [1]. Since the expiration of the patent of FDM in 2009, it has become one of the most common and affordable 3-D printing technology. As FDM has a relatively simple set-up structure and can uses relatively low-cost plastics such as polylactic acid (PLA), ABS and

polystyrene, it has been the most popular choice of 3-D printing for a wide variety of consumers ranging from hobbyists and academia to commercial industries

FDM uses a continuous thermoplastic polymer filament as a building material. The printer's extruder gear feeds the filament into the heater block to be melted and extruded out through its heated nozzle, as shown in Figure 2.1. Depending on the build material, the heater block is heated from 190°C up to 500°C. In some cases, the print bed is heated as well to avoid delamination and/or warping of the printed structure.

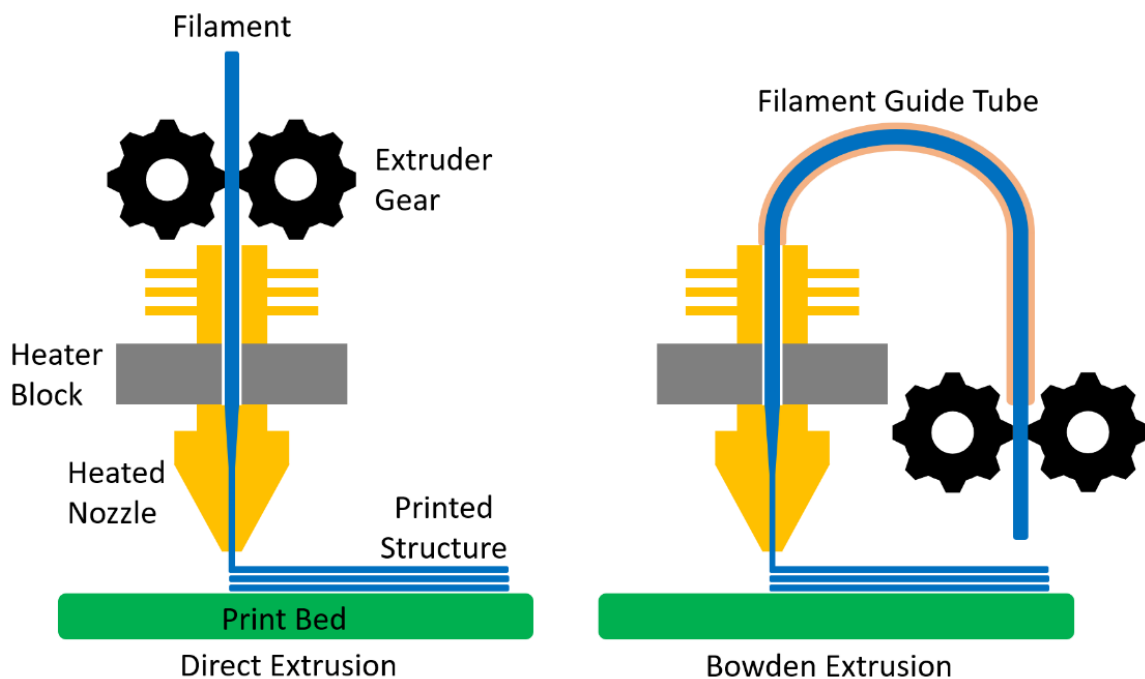


Figure 2.1: Illustration of FDM extruder structure.

Depending on the relative location of extruder gear, there are two types of FDM 3-D printers: direct extrusion; and Bowden extrusion. As shown on the left-hand side of Figure 2.1, the direct extrusion system has filament extruder gear positioned right above the heater block. Due to the relatively short distance between the extruder gear and the heated nozzle, it provides highly efficient power transfer to the filament hence resulting in better extrusion. With this type of extruder, relatively flexible filaments such as thermoplastic polyurethane (TPU) and nylon can be used do as there is not enough space for the filament to flex. However, because of the added weight of the whole extruder assembly

(including stepper motor and gears), it requires a stronger and tougher drivetrain for moving the assembly. Also, due to minimized space between extruder gear and the heater block assembly, it requires more complex maintenance.

Another type of extrusion system used by FDM is called Bowden extrusion. As shown on the right-hand side of Figure 2.1, the extruder gear is mounted far from the heated nozzle. The filament is guided from the extruder gear to the heater block using a flexible filament guide tube (often made of PTFE). This type of assembly significantly reduces the weight of the moving mass. Therefore, it has advantages such as fast print speed and less vibration. However, due to added distance between the printhead and the extrusion gear, it requires a much more powerful extrusion motor. Also, due to lag and flex between the gear and the printhead, in most cases, the Bowden extruder is not suitable for printing with relatively flexible filament.

There are many advantages of using FDM for rapid prototyping. Due to its simple structure, it can produce relatively large structures. As it is highly scalable, end users can relatively easily modify or build their own systems to accommodate their requirements. Another major advantage of using FDM comes from material flexibility. In principle, any material that has a melting point can be used for this technology. Basic low-cost polymer materials such as PLA and ABS are widely available to end-users. Moreover, more specialized materials such as conductive composite polymers and fiber reinforced polymers are available at relatively low-cost. Finally, unlike most other additive manufacturing techniques, parts printed with an FDM 3-D printer does not require complicated and potentially hazardous post-processing. Although sometimes it requires the physical removal of support structures, it is only a simple physical snap-off process.

In contrast, there is one major downside of using FDM 3-D printing technology. Inherently, extruded molten plastic cannot have infinitesimally thin layers. Using a smaller nozzle and higher print temperature can improve layer resolution. However, these adjustment causes other problems such as nozzle blockage and part warpage. Therefore, in most cases, the minimum layer height

that can be achieved reliably and repeatably is approximately 0.1 mm. Also, due to limitations on the nozzle's extrusion hole size, achieving a feature size of sub-0.5 mm is difficult. Finally, as extruded plastic is in a cylindrical shape, 100% infill is nearly impossible to achieve. Most of the time, the final FDM printed parts exhibit inherent inhomogeneity.

2.2 Vat Photopolymerization

Vat photopolymerization is another type of 3-D printing technology that has gained popularity over the past few years. The first patent of vat photopolymerization printing was filed in 1984, describing the stereolithographic apparatus (SLA) technique [2]. Vat photopolymerization 3-D printing uses UV curable photopolymer resin as a building material. External light sources, such as laser, UV lamp or projector, cure the resin layer by layer to build a final structure. Originally, the SLA process used a UV laser, possibly having a network of optical mirrors and lenses, to draw patterns on top or bottom surfaces of UV resin. More recently, rather than using a complicated laser-drawing mechanism, simpler and more affordable methodologies are being used. One example is masked stereolithography (MSLA); it uses simple UV light sources such as UV bulbs or LED with LCD screens to mask the required patterns. Figure 2.2 illustrates the working mechanism of SLA and MSLA techniques.

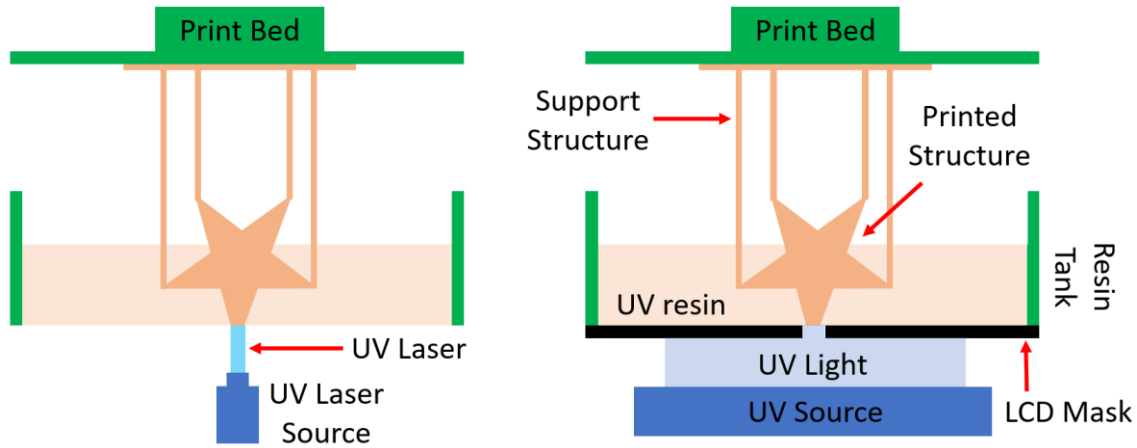


Figure 2.2 : Illustration showing Vat Photopolymerization printer mechanism: SLA (left); and MSLA (right).

Both SLA and MSLA 3-D printing offers a major advantage over FDM. It has much higher accuracy and resolution. In addition, the surface finish of the final structure can be almost as good as conventionally injection molded polymer parts. It is capable of printing extremely small features depending on the resolution. Moreover, as overall printing time is mostly dependent on the building height, a rapid mass-fabrication is possible. Unlike FDM technology, vat photopolymerization inherently produces homogeneous parts. Although SLA and MSLA printers share many similarities, for optimal results, appropriate printers need to be used depending on factors such as surface curvature, sizes, and required surface smoothness.

The original SLA 3-D printing uses a UV laser to draw patterns for a build layer, as illustrated in the right hand of Figure 2.3. The spot size of the UV laser decides the minimum feature size. In addition, due to relatively smooth laser movement, it tends to be excellent at producing curved or rounded surfaces [3]. Despite its ability to produce a smooth surface, when the parts have features or corners with sharp edges, SLA would not be the optimal choice. The SLA 3-D printing is ideal for printing parts that have curved features that require smooth surface finishes, such as mirrors and lenses.

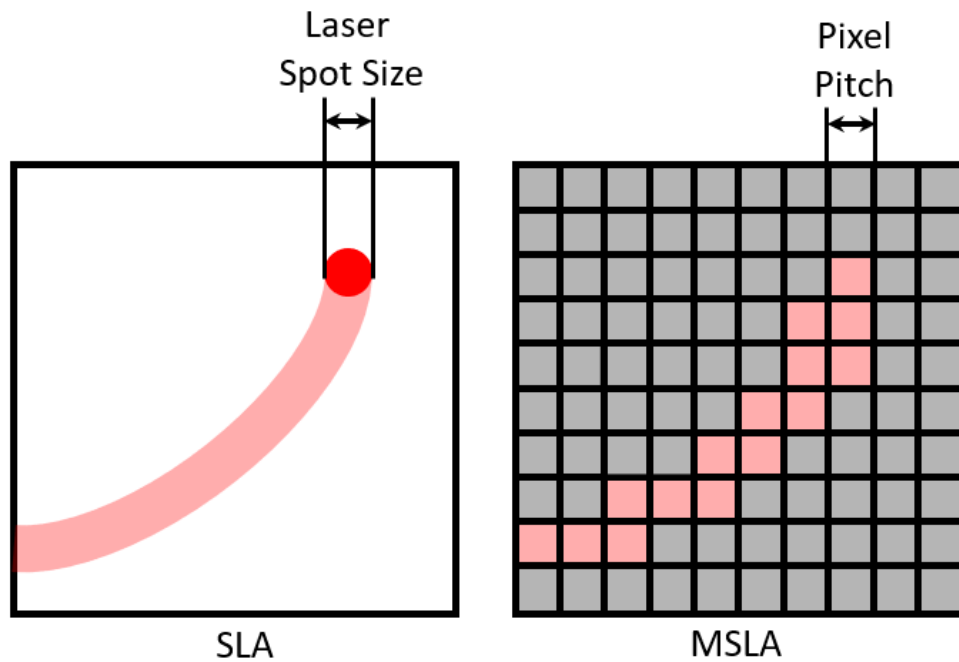


Figure 2.3: UV exposure mechanism of SLA (right) and MSLA (left)

MSLA 3-D printing employs an LCD masking screen that has a specific pixel pitch. The resolution (pixel pitch size) of LCD decides the minimum feature size. As shown on the left-hand side of Figure 2.3, curved or rounded surfaces will be pixelated, hence causing a rough surface. Despite this issue, MSLA excels at producing parts with sharp corners and edges. This makes the MSLA technology ideal for printing components such as rectangular waveguide filters and horn antennas. Moreover, due to rapidly improving display technologies, it is expected that more and more affordable MSLA printers with higher resolution will be available to end-users.

2.3 Other 3-D Printing Technologies

Other than FDM and Vat photopolymerization, there are several more types of polymer-based 3-D printing technologies that are less widely used at consumer-level applications. Here, two other types of 3-D printers that are most relevant for microwave and mm-wave applications will be discussed: material jetting and selective laser sintering (SLS).

Like SLA or MSLA, material jetting uses UV curable photopolymer as its primary build material. However, instead of selectively curing the UV resin directly inside the vat resin tank, the material jetting printer sprays micro-droplets of the resin through its nozzles, as shown in Figure 2.4. Then the UV light source attached to the printhead cures the projected droplets of the resin for a build layer. In most cases, wax-like support material will also be laid on the same layer, which can be removed through post-processing once the printing is finished.

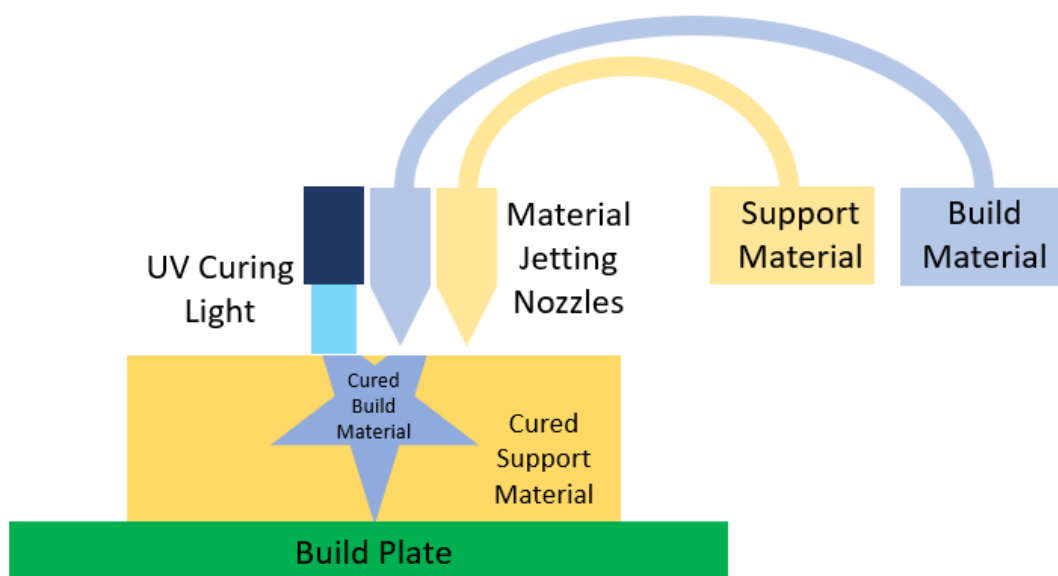


Figure 2.4: Illustration of material jetting 3-D printing

The printing mechanism of material jetting is similar to the conventional inkjet 2-D printer. Therefore, it can have multiple nozzles to have different types of colors of build material. Also, like the conventional inkjet 2-D printer, it has the line-by-line mechanism of scanning the whole build plate extremely fast. This allows the build speed to be fast and dimensional accuracy to be very good. Due to its capability of having separate support material that does not chemically bond to the cured build material, it can handle extremely complex structures without sacrificing surface smoothness.

Despite the advantages mentioned above of using the material jetting printer, it is not as widely used as FDM or Vat photopolymerization printers because of several factors. Due to the complexity of the material droplet jetting mechanism, it is orders of magnitudes more expensive to purchase, run and maintain than the other 3-D printers. Moreover, as it requires very specific material, not many material selections are available currently.

Another 3-D printer that is relatively widely used is selective laser sintering (SLS). Polymer SLS uses a high-powered laser to sinter powdered polymer and builds a layer of the required structure, as shown in Figure 2.5. When the sintering of one layer is finished, the build plate moves down by single layer height, and the polymer powder spreader fills and spreads fresh polymer powder on top of the previous build layer. As the building structure is suspended in the powder of un-sintered polymer powder, it does not require a separate support structure. Also, depending on the material, as it utilizes a high-powered laser, sintering time can be minimized hence resulting in extremely fast overall printing time.

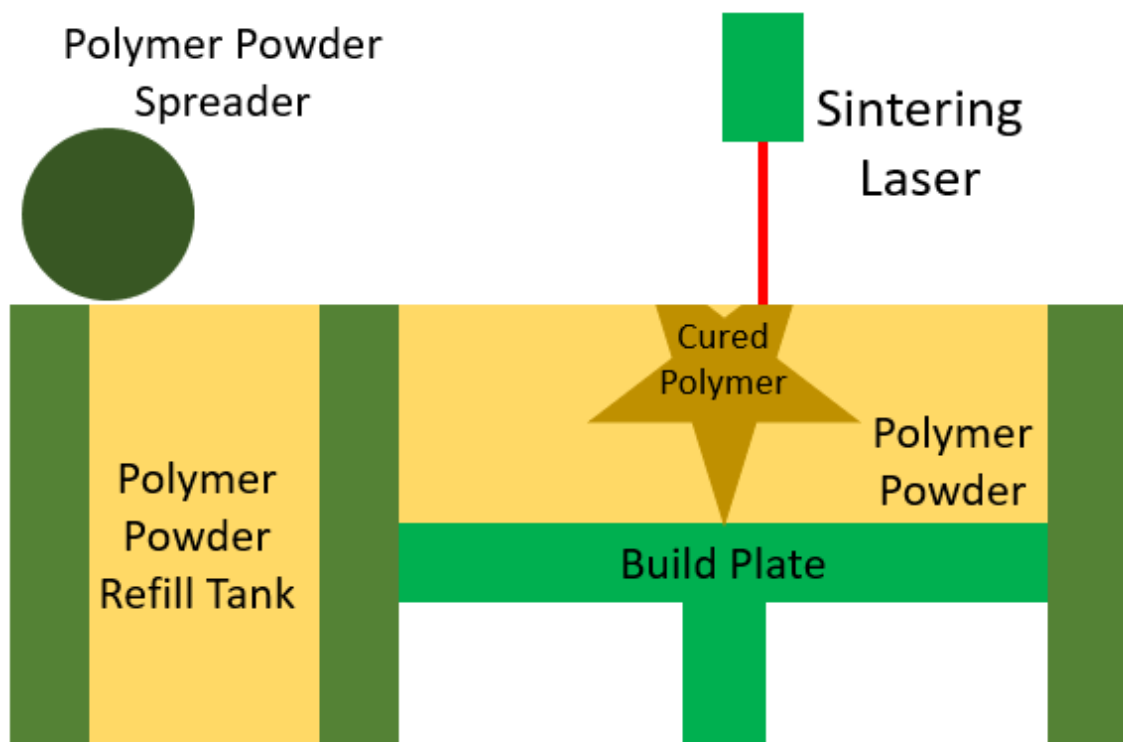


Figure 2.5: Illustration of polymer-based SLS 3-D printing

However, due to the nature of the powdered polymer, the finished product is inherently porous and has rough surfaces. As the build material is sintered at high temperature and gets cooled immediately, internal thermal stress causes the printed structure to warp and shrink significantly. In addition, because the raw build material is a fine powder of various polymer types, it requires extra safety precautions. Due to the respiratory hazard that the polymer powder causes, it requires filtered air ventilation and a special sealed chamber for post-processing.

2.4 3-D Printer Comparison and Scoring

As discussed in previous subsections, each 3-D printing technology has its own unique advantages and disadvantages. Factors such as material availability, print resolution, surface quality, build size, post-processing and cost, allows end-users to choose an appropriate 3-D printing technology for their applications. Table 2.1 shows the comparison and relative scoring of the five printing technologies covered in this section: FDM, SLA, MSLA, material jetting and SLS.

Table 2.1: 3-D printing technology comparison and scoring

	FDM	SLA	MSLA	Material Jetting	SLS
Material	5	4	4	2	3
Resolution	1	5	5	4	3
Surface Quality	1	5	4	5	2
Build Size	5	3	2	4	2
Ease of Post Processing	5	3	3	2	1
Cost	5	3	5	1	2
	Poor	Average			Excellent
Rating Scale	1	2	3	4	5

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3 3-D Printing and Post-Processing

To 3-D print a component, a 3-D model needs to be generated first. Most of the conventional computer-aided design (CAD) tools are appropriate for the process. Once modelling is finished, the 3-D models need to be converted into STL mesh file format. The STL file converts 3-D surfaces into triangulated approximation. For microwave and mm-wave applications, as surface geometry is critical for the functionality of the component, it is important to use the finest triangular mesh setting.

For FDM 3-D printers, the prepared STL file needs to be converted into a G-code file. G-code is one of the most common computer numerical control (CNC) programming languages to be used for FDM 3-D printing. It converts 3-D geometry to a series of commands that controls actuators such as servos or stepper motors for moving a 3-D printer head to the required coordinate. In most cases, commercially available FDM printers come with their own STL-to-G-code conversion software that is optimized for specific hardware. It is also possible to generate a manually customized G-code if required. Finally, the generated G-code needs to be uploaded into an FDM printer, and the 3-D printing process will begin.

For SLA and MSLA printers, the prepared STL file need to be converted into multiple image files. Each image file corresponds to a single build layer and controls UV exposure by either masking (MSLA) or drawing (SLA). Conversion can be done through slicing software that comes with the 3-D printer. The slicer software can then be used to adjust the UV exposure settings.

Unlike conventional precision machine-based fabrication processes, consumer-level 3-D printers such as FDM and SLA does not always produce perfect dimensional accuracy. Factors such as print orientation, layer height, and post-processing techniques significantly affect dimensional accuracy. Therefore, several iterations of fabrication and dimensional measurements might be necessary until achieving desired performances. The 3-D printed components described in this thesis followed standardized processes as described in this

chapter, but sometimes several versions were fabricated until reaching acceptable measured performances.

3.1 Print Orientation and Layer Height

For both FDM and SLA/MSLA, print orientation plays a vital role in terms of surface quality and dimensional accuracy. Print orientation can be adjusted using the G-code generation software. Inherently, 3-D printed components can have visible layer lines along the build direction. In addition, as 3-D printers always require a build surface to add a layer of material, the components often require additional support structure to prevent overhang. Surfaces that have support structure often have rough surfaces and dimensional inaccuracies due to leftover material from post-processing. Therefore, it is important to make sure that the components are orientated such that there is no support structure attached to important surfaces.

3-D printers can only achieve finite layer thickness. Therefore, surfaces that have curved features relatively normal to the build direction can never be perfectly smooth. One way of minimizing this problem is by decreasing the layer height. However, most of the consumer level FDM and SLA/MSLA printers are only capable of producing layer thicknesses larger than 0.1 mm and 0.02 mm, respectively. Another way of minimizing this problem is by orientating the printed components in certain ways. Inherently, due to its layer-by-layer building mechanism, a 3-D printer cannot replicate a perfectly smooth surface. Therefore, parts with curved surfaces require build orientation to be optimized. As shown in Figure 3.1(a), aligning the curved surface relatively parallel to the build direction (left) has a much better discretized representation of the curvature than the one aligned relatively normal to the build direction (right). In addition, as shown in Figure 3.1(b), by changing the orientation carefully, the support structure can be significantly reduced. This will allow much better surface quality, faster post-processing and reduced material waste.

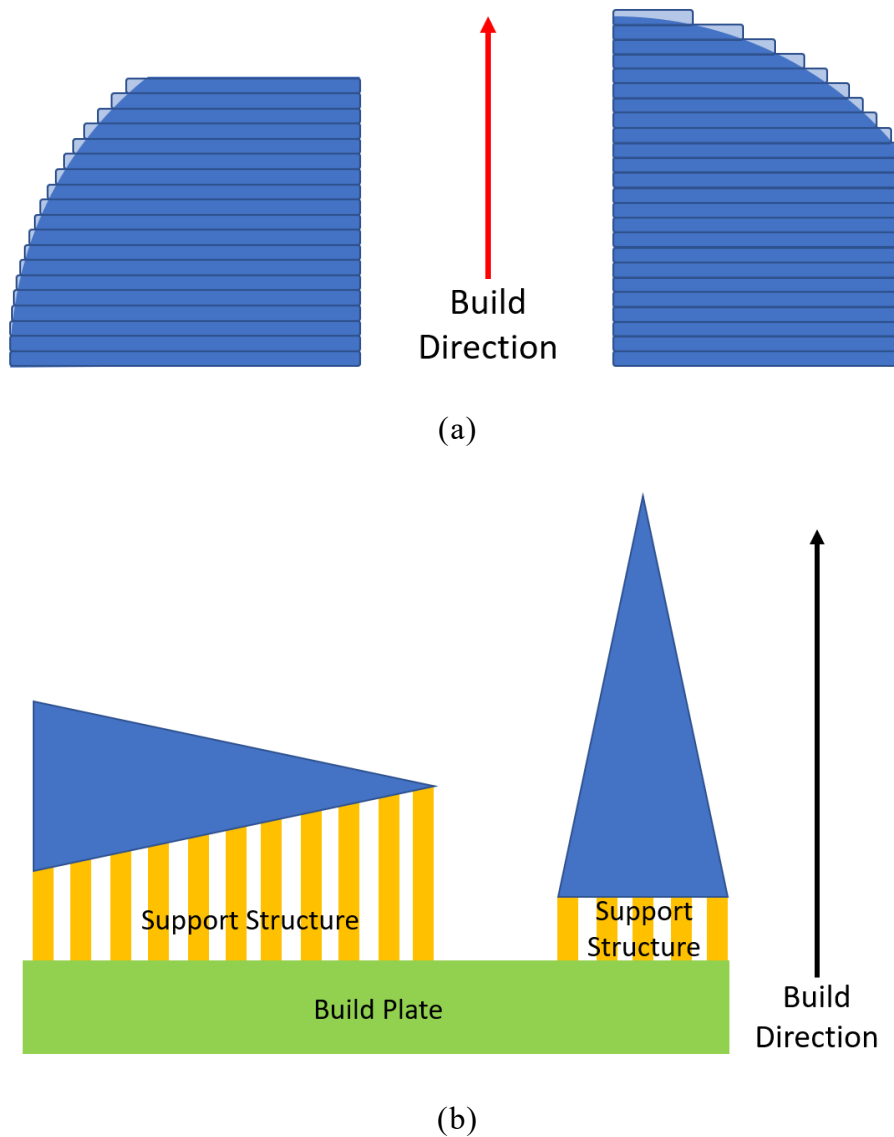


Figure 3.1: Illustration of effect on changing build direction

For mm-wave quasi-optical components such as parabolic mirrors and dielectric lenses, to have the best result, a combination of optimum print orientation and minimum layer height is essential. 3-D printed parts need to be orientated to locate the support structure away from the most important surfaces, such as the reflecting faces of the mirrors and curved surfaces for lenses. Also, minimizing the layer thickness while keeping the curved surfaces as parallel as possible to the build direction were found to give the most accurate and smoothest results. It is important to note that several iterations of fabrication for each component are necessary for the best result.

3.2 3-D Printer Settings

In this study, FDM was used mainly for building housings, brackets and adapters for mm-wave and quasi-optical applications. Also, as FDM provides a diverse selection of build material with various dielectric properties, it was used for printing components such as dielectric phase shifter and radiation absorbent material (RAM). Depending on the requirement or material properties of the components, optimal printing parameters were found and applied. As all 3-D printers have to be set-up differently for achieving an optimal result, in this subsection, general rules and methodologies of choosing the right parameters will be discussed.

The most important parameter for FDM printing is the heated nozzle temperature. Depending on the build material, printing temperature needs to be tuned carefully in order to produce the best result. When the printing temperature is too low, the material's viscosity might not be sufficiently low. This prevents enough material from being extruded and causes problems such as part warpage. When the printing temperature is too high, too much material can be extruded and cause surface defects. Also, to prevent large components from warping during the printing process, the build plate needs to be heated to make sure the extruded plastic to be cooled evenly across a layer at a controlled rate. In some cases, a heated enclosure could be necessary to prevent excessive warping.

Components that require accurate geometrical features and smooth surfaces such as dielectric lenses, parabolic mirrors and horn antennas were fabricated using SLA and/or MSLA printers. Therefore, having the highest resolution and minimizing part warpage is crucial. UV exposure settings need to be adjusted carefully for different layer height settings in order to prevent parts from warping and/or failing. Smaller layer thickness requires a shorter UV exposure time. However, insufficient exposure time cause problems such as layer delamination. As with FDM, each resin based 3-D printer have a different optimal set-up.

Specific FDM, SLA and MSLA settings used for each component will be discussed in the following chapters.

3.3 Post-Processing

Most of the 3-D printing processes, including FDM, SLA and MSLA, require different levels of post-processing. For FDM, simply removing support structures using tools such as pliers or knives is usually sufficient. Occasionally, sanding might be required for applications that require smooth surfaces. However, as sanding affects the physical dimensions of 3-D printed parts, it is not recommended. Resin-based 3-D printing almost always needs post-processing that involves both physical and chemical removal of unwanted structures and uncured resin residue. In addition, post UV curing of cleaned parts is often necessary. In this subsection, detailed SLA and MSLA post-processing techniques that were developed for fabricating microwave and mm-wave components will be discussed.

When SLA or MSLA printers finish printing processes, because the components are printed while submerged inside their resin tanks, excess uncured resin has to be cleaned. Also, due to reduced dimensional accuracy for layers close to the build plate, the components are often suspended with tree-like support structures that need to be physically removed. Depending on the type of UV resin, post-processing techniques differ. Two types of UV resin were used for fabricating the components that will be discussed in a later section; conventional UV photopolymer resin and water washable resin require IPA/ethanol and water, respectively, for the cleaning process. Table 3.1 shows the step-by-step procedures that were developed to be suitable for fabricating the microwave and mm-wave components described in later sections.

Table 3.1: Post-processing procedures for SLA and/or MSLA 3-D printed components.

Process	Conventional Resin	Water Washable Resin
1. Pre-Rinsing	Room temperature IPA bath with hand agitation	Room temperature water bath with hand agitation
2. Support Structure Removal	Mechanical removal of support structures using pliers	
3. Detailed Cleaning	- IPA jet cleaning - Hand agitation in IPA bath up to 10 minutes	- Water jet cleaning - Ultrasonic bath up to 30 minutes
4. Rinsing	Rinsing using IPA squeeze bottle	Rinsing under running water
5. Drying	Drying most of the liquid off, using compressed air	
6. Post Curing 1	Up to 10 minutes under UV lamp (rotation required for curing all surfaces)	
7. Post Curing 2	Slow curing under natural sunlight do prevent warping	

There are several steps of repeated cleaning and rinsing as insufficient cleaning causes the residue to cure and harden during the post-curing processes hence resulting in dimensional inaccuracies and surface defects. Also, two different steps of the post-curing process were found to give the best result. Using a UV lamp cures the surfaces quickly and prevents surface defects as it minimizes dust build-up on uncured surfaces. However, it was found that a relatively strong and fast curing process causes parts to warp and deform. Therefore, a two-step post-curing process was adopted.

3.4 3-D Printing Safety

Polymer-based 3-D printing allows end-users freedom in design, fast, easy and cost-effective ways of prototyping and small-batch production. However, as the technologies are relatively new, appropriate facilities for safe fabrication and post-processing are not commonly available for academic and home-based users. For example, Vat photopolymerization printers require potentially hazardous and toxic materials such as UV resin and organic solvent. In this subsection, appropriate lab-based, personal protective equipment (PPE) for safe FDM and Vat photopolymerization 3-D printing will be discussed. In addition, an example will be given for setting-up a safe lab environment for selected types of 3-D printing techniques.

As discussed in previous sections, FDM is the most common 3-D printing technique. It is widely used in an environment such as school classrooms and university labs. As FDM utilizes molten plastic as its primary build material, it generates fumes and micro-particles that can be potentially hazardous. Studies have shown that common FDM 3-D printing material such as ABS and PLA produces many different types of hazardous volatile organic compound (VOC) such as styrene, betadine, and formaldehyde [1]. Moreover, as a countless number of companies produce FDM filaments with unclear substances, potential danger can be even more serious. Therefore, long term exposure to micro-particles and volatile organic compound emissions needs to be prevented [2], [3]. Another potential danger of FDM comes from heated nozzles. During the printing process, the 3-D printer's heated nozzle reach more than 200°C, which can cause a potential burn hazard.

For the Vat photopolymerization printing technique, the potential health hazard is mainly caused by the UV photopolymer resin base material. Generally, uncured UV resin requires some chemical safety measures. Uncured resin can cause skin and eye irritation as well as respiratory problems [4-6]. It is also known to cause nausea and headache when high vapor concentration is inhaled [6]. In addition, as SLA printer uses Class 1 UV laser, appropriate precaution and education are required for the users. For post-processing parts printed with Vat

photopolymerization 3-D printer, different types of an organic solvent such as isopropanol (IPA) and acetone are used. Therefore, it is a potential fire hazard and can cause skin and respiratory irritations.

The problems associated with micro-particles, fume and VOC can be easily rectified using an enclosure with filtered ventilation. Using an enclosure with HEPA filter and active charcoal filter reduces particles and VOC emissions significantly [7]. Therefore, using a commercially available fume hood or fume extractor equipped with HEPA and an active carbon filter is recommended. Properly enclosing the 3-D printers also prevents users from accidentally touching heated nozzles. Also, using appropriate PPEs such as lab coats, nitrile gloves and goggles provide enough protection for using potentially hazardous 3-D printing and post-processing materials. For the printers such as SLA, which has Class 1 lasers, it is important for users to get an appropriate education for laser safety. However, as most of these printers has laser beam paths enclosed within the printer itself, as long as the users follow the manufacturer's instructions, any kind of exposure is unlikely. Nevertheless, proper PPEs such as UV laser safety goggles should be prepared for unexpected events. An example of a safe lab set-up for 3-D printing will be discussed in the following subsection.

3.4.1 Example of Safe 3-D Printing and Post-Processing Set-up

For fabricating most of the devices in the following sections, a combination of consumer-level FDM, SLA and MSLA 3-D printers are used. This involved several hours per day of 3-D printing and post-processing. In order to make sure that the fabrications processes were safely executed in a university-level laboratory that did not have adequate equipment and infrastructure for safe 3-D printing, customized lab set-up and certification was necessary.

To address issues associated with fume generation and ventilation, either a commercial fume cupboard or custom-made fume extraction enclosure was required. To accommodate a large format FDM printer (Raise 3D Pro 2 [8]) and an SLA printer (Form 3 [9]) having an external dimension of 620 mm × 590 mm × 760 mm and of 405 mm × 375 mm × 530 mm, respectively, a bespoke fume extraction enclosure was designed and built.

Working with Imperial College's local exhaust ventilation team, a bespoke fume extraction enclosure was built. The enclosure is built to have all the necessary components such as baffles, sash door, fume extractor, inlet and exhaust outlet [10]. To ensure sufficient airflow and effective filtering, a commercial fume extractor (BOFA AD Oracle iQ [11]) with pre, HEPA and active carbon filter was selected. The illustration of design and CAD drawing is shown in Figure 3.2.

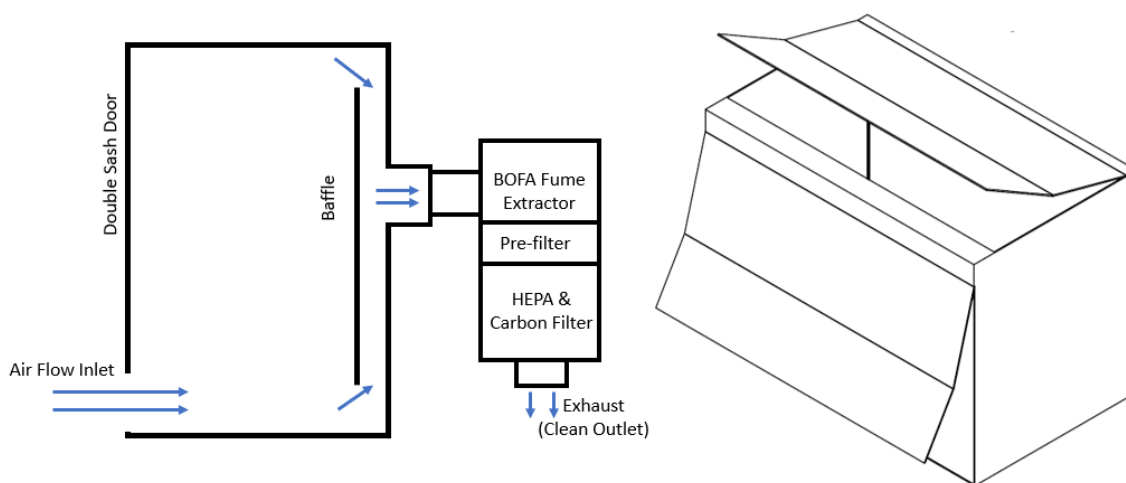


Figure 3.2: Fume extraction enclosure: element illustration (left); and CAD drawing (right)

For building the enclosure, 30 mm $\times$ 30 mm aluminum extrusions were used as a frame, and 6 mm thick clear acrylic panels were used as outer shells. The final structure was thoroughly tested and certified by Imperial College's health and safety team to be used for 3-D printing and post-processing activities. It is important to note that this kind of custom-built fume extraction system requires regular maintenance and certification. Also, for smaller MSLA 3-D printers and general handling of organic solvents and UV resins, a standard commercial fume cupboard with pre-, HEPA and active carbon filter was purchased. For safe storage of all the chemicals, a flame-safe chemical cabinet was used. For minimizing potential hazards due to transportation of chemicals and parts, the bespoke enclosure, commercial fume cupboard and cabinet were located right next to each other. A photograph of the final lab set-up is shown in Figure 3.3.

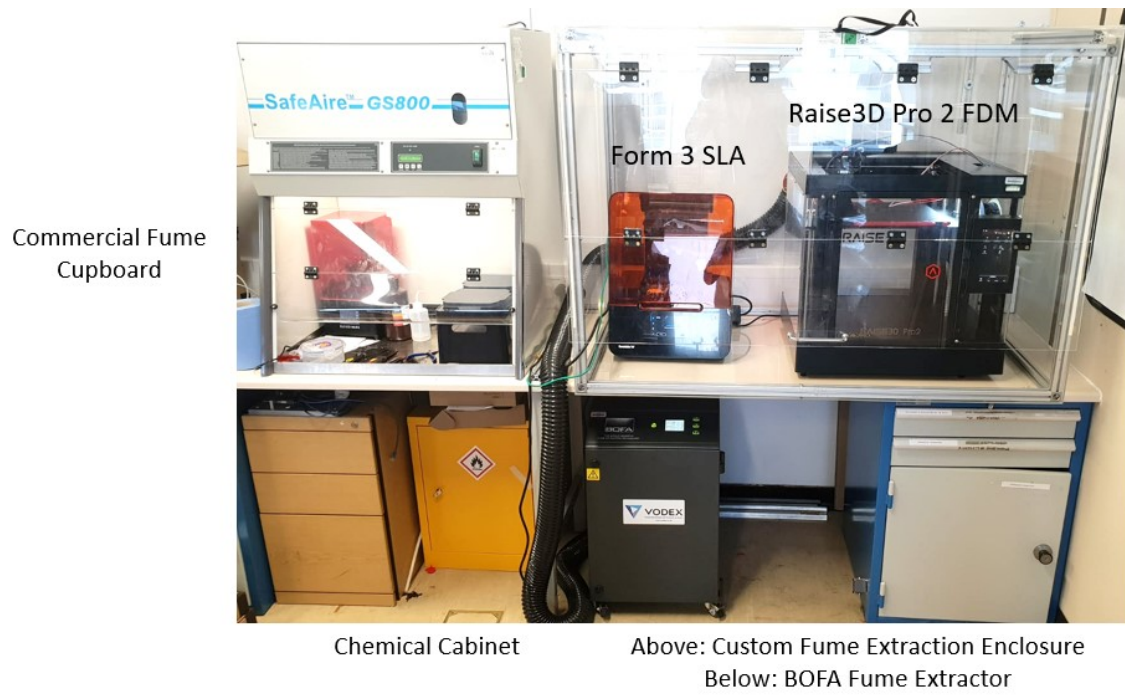


Figure 3.3: Photograph of 3-D printing and post-processing lab set-up

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4 Ku-band Steerable Phased-Array Antenna

4.1 Introduction

The content in this chapter is based on previously published work by Shin *et al.* [1]. Phased-array antennas are widely used for many applications, such as in modern radar and space communications systems. Phased-array antennas have many identical stationary antenna elements that form a steerable beam. It requires multiple sets of highly customized components, such as transmission lines or guided-wave structures, power splitters, phase shifters, feed networks and radiating elements. Due to the nature of phased array antenna's geometrical complexities, with conventional fabrication techniques, the parts need to be fabricated separately and assembled. As a result, it requires many waveguide flange connections; each contacting pair introducing a degradation in performance. In contrast, 3-D printing technologies allow more complex 3-D structures to be manufactured. Also, the number of parts can be significantly reduced with 3-D printing, minimizing the number of flanges. Moreover, with unnecessary flanges removed, the overall weight can be significantly reduced as well.

This work describes in detail the design, manufacture and measured results of a fully 3-D printed 4-element steerable phased-array antenna subsystem. Here, 26 circuit elements ($1 \times$ flange, $6 \times$ mitred bends, $3 \times$ power splitters, $4 \times$ tunable delay lines, $4 \times 90^\circ$ waveguide twists, $4 \times$ bent interconnects and $4 \times$ H-plane sectoral horn antennas) are manufactured as a fully integrated, complete subsystem. In addition, fully 3-D printed variable-position dielectric insertion mechanisms have been developed to implement tunable delay lines. All the measurements are performed by the UK's National Physical Laboratory, with reference to the UK's primary national measurement standards.

4.2 Design

The beamforming network and individual horn antennas have been designed as a fully-integrated phased-array subsystem. The basic block diagram of the beam-forming network and antenna array is shown in Figure 4.1. The subsystem consists of 7 different types of MPRWG components: flange, bends, power splitters, tunable delay lines, twists, interconnects and horn antennas. All the waveguide components are based on conventional WR-62 waveguide, operating at Ku-band, having internal cross-sectional dimensions of $a = 15.799$ mm and $b = 7.899$ mm. The design has been optimized for operation between 15 and 17 GHz.

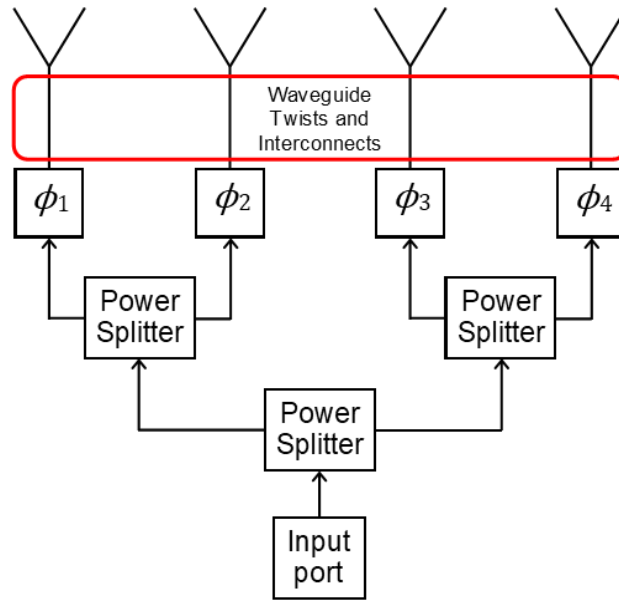


Figure 4.1: Basic block diagram of the 4-element phased-array antenna subsystem.

With the 4-element phased-array antenna having a single input waveguide port, three symmetrical power splitters were introduced to create a 4-way corporate feed. Ideally, the conventional 3 dB power splitter design employs a thin metallic septum to achieve optimal performance. However, a 3-D printed thin vertical polymer wall is not structurally stable. Moreover, it may exhibit warping during the electroplating processes. Therefore, the septum was replaced by a 0.85 mm diameter cylindrical stainless-steel pin, with the use of

electromagnetic software simulations to determine the optimal location. The two designs for the 3 dB power splitter are illustrated in Figure 4.2.

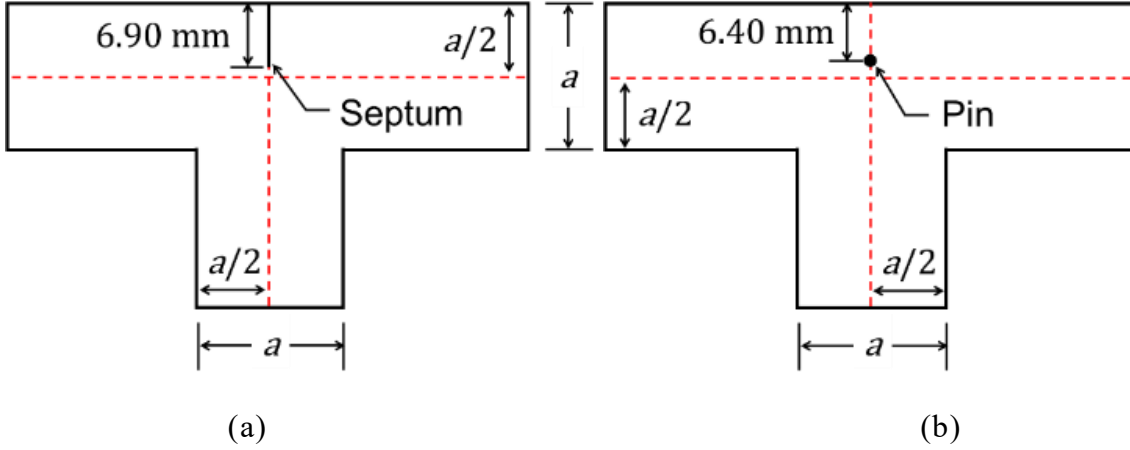


Figure 4.2: Illustration of 3 dB power splitter: (a) conventional design using a septum; and (b) pin solution.

A phase shifter is a control device found in many microwave communications, radar and measurement systems. There are subtle differences between the two main generic types of phase shifter; true phase shifter and true delay line [2]. True delay lines find many applications in general wideband microwave signal processing applications, including beam-forming networks for phased-array antennas [2].

With phased-array antennas, phase steering and time steering are implemented by employing true phase shifters and true delay lines, respectively, as defined by the following:

True phase steering:

$$\Delta\phi = \Delta\phi(f) \text{ and } \theta(f) = \sin^{-1} \left\{ \frac{\lambda}{d} \cdot \frac{\Delta\phi}{360} \right\} \quad (4.1)$$

True time steering:

$$\Delta\phi(f) = \frac{f}{f_o} \Delta\phi(f_o) \text{ and } \theta = \sin^{-1} \left\{ \frac{\lambda}{d} \cdot \frac{\Delta\phi(f)}{360} \right\} \quad (4.2)$$

where, $\Delta\phi = \phi_i - \phi_{i-1}$ represent the difference in phase between the i^{th} and $i^{th} - 1$ antenna elements, θ is the steering angle from broadside boresight, f is the frequency, λ is the corresponding wavelength in free-space and subscript ‘o’ refers to the center frequency of the band of operation.

A 3-D printed MPRWG variable phase shifter has been developed that exploits a moveable dielectric flap insert, operating across X-band [3]. Its frequency behaviour resembles a true phase shifter in the lower third of the band and a true delay line in the upper third of the band, with an undefined characteristic mid-band.

With the Ku-band design adopted in this work, the dielectric insert taper profile and insertion mechanism were completely redesigned to improve RF performance, structural stability and control. The dielectric insert was designed to be FDM 3-D printed using ABS, with Formfutura’s Premium ABS (Frosty White), having X-band values of dielectric constant $\epsilon_r = 2.31$ and loss tangent $\tan\delta = 20 \times 10^{-4}$ [4]. The guided wavelength,

$$\lambda_g = \frac{\lambda}{\sqrt{\epsilon_{reff} - \left(\frac{\lambda}{2a}\right)^2}} \quad (4.3)$$

is dependent on the effective relative permittivity within the section of waveguide containing the dielectric insert ϵ_{reff} (with $1 < \epsilon_{reff} < \epsilon_r$). When the dielectric insert penetrates the region of maximum electric field for the dominant TE₁₀ mode, with more fractional volume filling the waveguide section, ϵ_{reff} increases and λ_g decreases [5]. As a result, the transmission phase within the fixed length of delay line increases.

To achieve the maximum relative phase shift and minimum losses, the dielectric inserts were designed to have a rectangular shape with elliptically tapered edges on both front and rear ends. Brackets were designed to have a 3-D printed calibrated screw mechanism that inserts and extracts the dielectric pieces linearly and evenly through the slots. A pair of guide rails were added to each

bracket to add extra stability. The CAD rendered drawing of a dielectric insert, bracket and screw mechanism are shown in Figure 4.3(a).

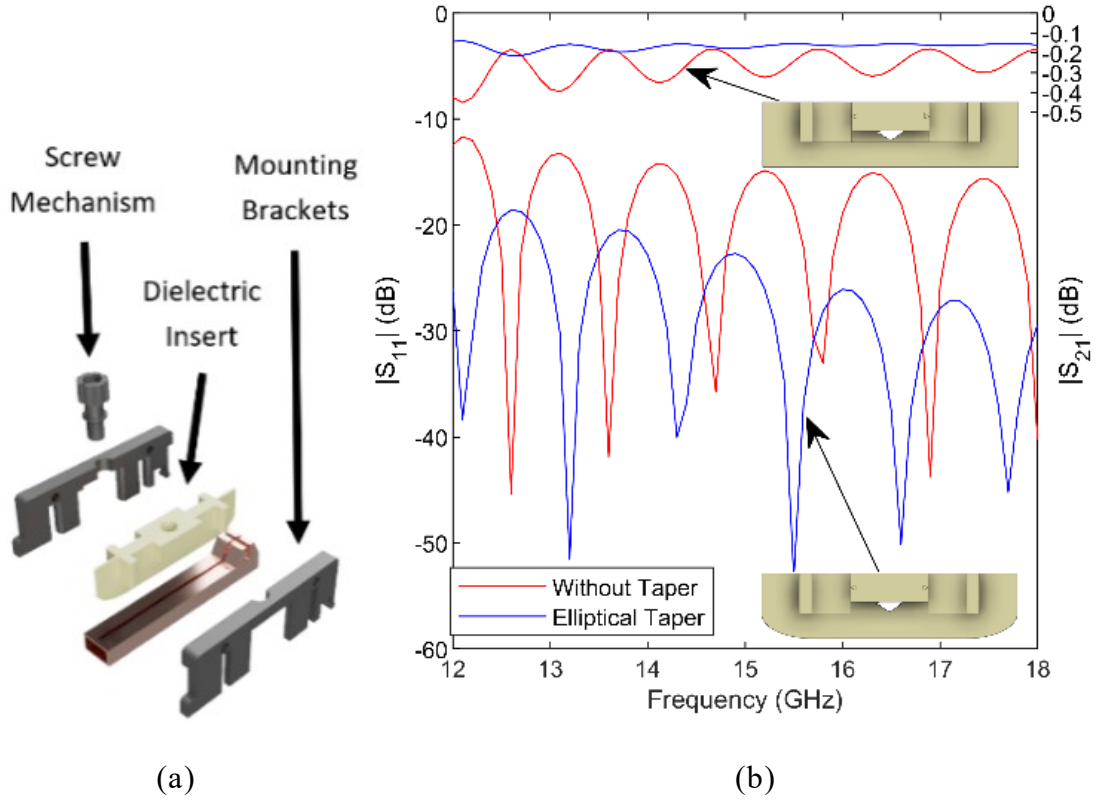


Figure 4.3 : Dielectric phase shifter: (a) Rendered CAD design of dielectric inserts and brackets; and (b) simulated insertion and return loss of phase shifter across Ku-band for 2 variations of dielectric insert shapes: without and with elliptical taper.

Electromagnetic (EM) simulations were performed on the delay line section using ANSYS HFSS to optimize the size and shape of the tapered edges of the dielectric inserts. Simulated scattering (S-)parameter results are shown in Figure 4.3(b), which clearly show that the elliptical taper shape has superior return and insertion loss behavior across Ku-band. With the dielectric insert at maximum depth, relative phase shifts of 262° and 288° are predicted at 15 and 17 GHz, respectively.

Figure 4.4 shows the predicted phase shifter performance for different levels of dielectric insert penetration depth g . The relative phase shift (when referenced to a non-penetrating dielectric insert) exhibits complicated features across the

waveguide band; approximating to a true phase shifter at the lower band edge frequency, a true delay line at the upper band edge frequency and an undefined mid-band region. This is because a MPRWG is inherently a very frequency dispersive transmission medium. This resulting frequency behavior can also be seen in the X-band proof of principle [3]. The two phase-shifting boundaries, seen in Figure 4.4(a), are defined here by a 1° deviation from either true phase shifter or true delay line characteristics, for the reference level having $g = 0$. It can be seen, between 15 GHz and 17 GHz, the phase shifter resembles a true delay line for low levels of relative phase shift, but the undefined region extends further up in frequency with higher levels of relative phase shift.

Figure 4.4(b) shows the corresponding differential-phase group delay frequency responses for the phase shifter between 15 GHz and 17 GHz. Clearly, the phase shifter exhibits true delay line characteristics in the region identified in Figure 4.4(b), with this approximation being poorer in the transition region.

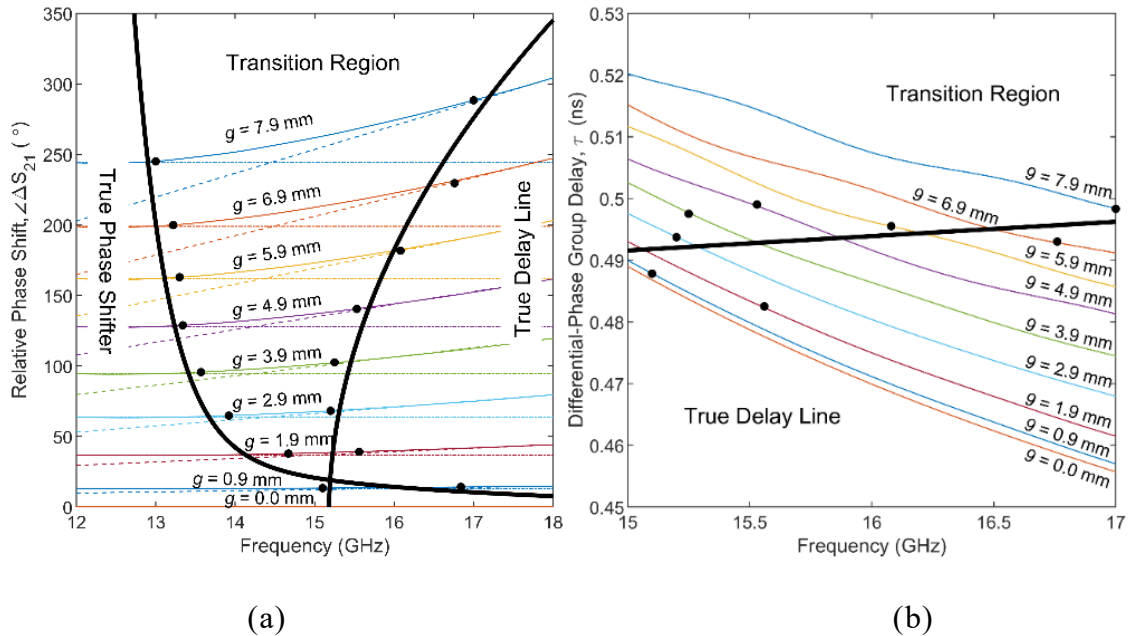


Figure 4.4: EM simulated phase shifter performance: (a) relative phase shift (when referenced to a non-penetrating dielectric insert having $g = 0$); and (b) differential-phase group delay response. Black dots represent the 1° deviation from either true phase shifter or true delay line characteristics and the solid black lines represent the corresponding curve fits.

The independent tunable phase shifters are followed by 90° twists, followed by bent waveguide interconnects required to feed the individual H-plane sectoral horn antennas, while maintaining equal delay paths when there is no steering from the broadside boresight. For optimal performance, the distance between each array element d is set to half the free-space wavelength $\lambda = 20$ mm at 15 GHz (i.e., $d = \lambda(15 \text{ GHz})/2 = 10$ mm). A WR-62 based H-plane sectoral horn antenna is ideal for achieving this, as its E-plane height dimension is sufficiently smaller than the operational free-space wavelength (i.e., $b < \lambda(15 \text{ GHz})/2$). The internal dimensions for the H-plane sectoral horn antenna design are shown in Figure 4.5.

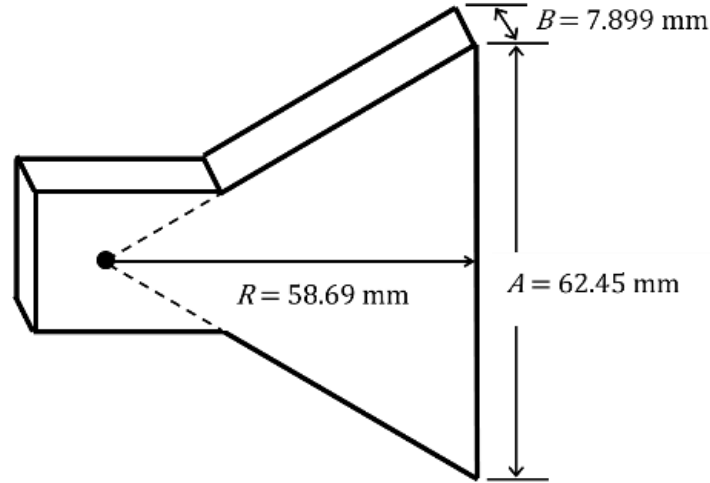


Figure 4.5: Illustration of the H-plane sectoral horn design.

With infinitesimally thin metal walls made with a perfect electrical conductor (PEC), the theoretical directivity of the single horn antenna D_e is [6], [7],

$$D_e(\text{dB}) = 10 \cdot \log_{10} \left\{ \varepsilon_t \varepsilon_{ph}^H \cdot \frac{4\pi}{\lambda^2} \cdot Ab \right\} = \begin{cases} 9.8 \text{ at } 15 \text{ GHz} \\ 10.5 \text{ at } 17 \text{ GHz} \end{cases} \quad (4.4)$$

where, the aperture taper efficiency factor $\varepsilon_t = 8/\pi^2$ and the phase efficiency factor is given by [6]:

$$\varepsilon_{ph}^H = \frac{\pi^2}{64\xi} \cdot \{[C(p_1) - C(p_2)]^2 + [S(p_1) - S(p_2)]^2\} \quad (4.5)$$

and parameters p_1 and p_2 are [7],

$$p_1, p_2 = \frac{1}{\sqrt{2}} \cdot \left[(8\xi)^{-\frac{1}{2}} \pm (8\xi)^{+\frac{1}{2}} \right] \quad (4.6)$$

and the Fresnel integrals, $S(x)$ and $C(x)$ are a function of the fractional aperture distribution phase error ξ ,

$$\xi = \frac{A^2}{8\lambda R} = \begin{cases} 0.42 & \text{at 15 GHz} \\ 0.47 & \text{at 17 GHz} \end{cases} \quad (4.7)$$

The complete phased-array antenna's broadside boresight directivity (D_A) was calculated using [6]:

$$D_A(dB) = D_e(dB) + D_i(dB) = \begin{cases} 15.8 & \text{at 15 GHz} \\ 17.1 & \text{at 17 GHz} \end{cases} \quad (4.8)$$

where, the directivity for a broadside linear array of N uniformly-excited isotopically radiating elements is given by,

$$D_i(dB) = 10 \cdot \log_{10} \left\{ 2N \frac{d}{\lambda_0} \right\} = \begin{cases} 6.0 & \text{at 15 GHz} \\ 6.6 & \text{at 17 GHz} \end{cases} \quad (4.9)$$

The electromagnetic CAD simulations assume PEC metallization (which represents a good approximation to copper at Ku-band), but with metal wall thicknesses for the horns corresponding to those in the actual prototype antenna array. This dramatically reduces radiation backlobe levels (by reducing the backward radiation leakage caused by knife-edge diffraction that results from infinitesimal thin walls) and, thus, increases predicted antenna gain at boresight beyond that calculated in (4.8).

CAD drawings for the main body of the phased-array antenna are shown in Figure 4.6. To validate the final design of the complete phased-array antenna subsystem, simulations were performed using HFSS with PEC waveguide walls (to minimize the computational resources needed for such a large and complicated structure). Figure 4.7 shows the simulated internal E-field patterns

at 15 GHz, with all the individual components functioning as expected. For example, it can be seen in Figure 4.7(b), when a dielectric insert has maximum penetration, it accommodates an extra $\sim 3\lambda_g/4$, when compared to zero penetration.

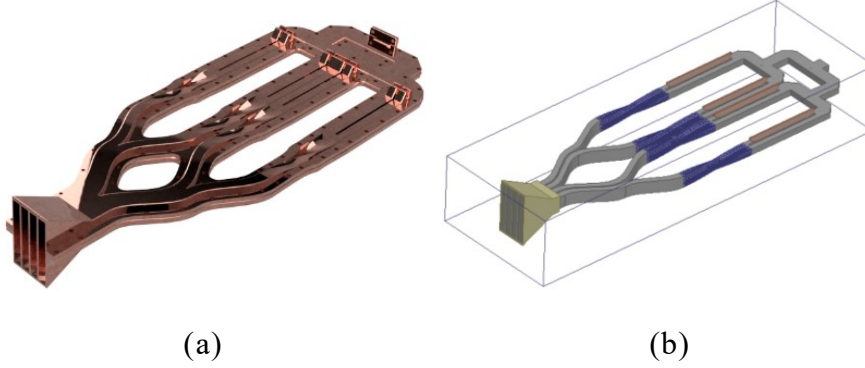


Figure 4.6: CAD design of the main body of the phased-array antenna subsystem: (a) rendered 3-D model; and (b) simulation model with radiation boundary.

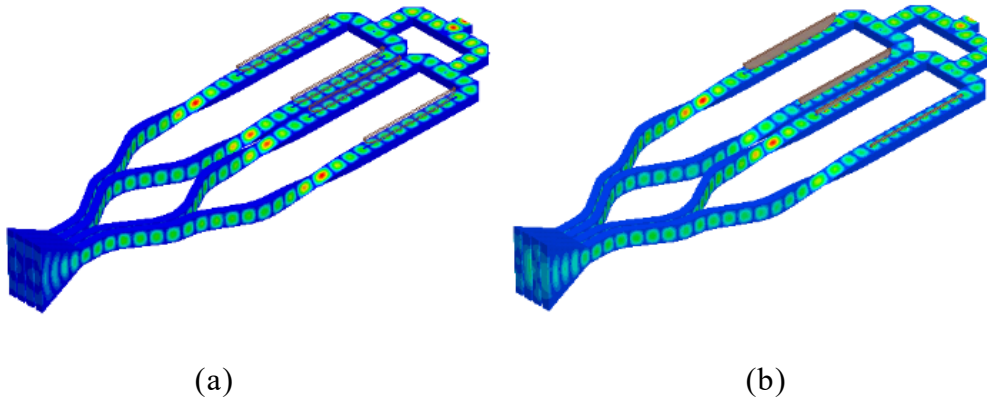


Figure 4.7: Simulated internal E-field patterns at 15 GHz: (a) with all the dielectric inserts fully inserted; and (b) with dielectric insertion setting for maximum beam steering angle.

4.3 Manufacture

To manufacture the complete phased-array antenna subsystem, two types of 3-D printing technologies were employed: (i) PolyJet, an industrial-level process with ultra-high precision; and (ii) low-cost, entry-level desktop FDM.

4.3.1 PolyJet

A PolyJet 3-D printer (Stratasys Objet500 Connex3) was used for fabricating the MPRWG parts of the phased-array antenna subsystem. This printer is capable of printing large models at high resolution (maximum build area of 490 mm x 390 mm x 200 mm, with a minimum layer height of 16 μm) by using its polymer jetting technology. The PolyJet printer jets photopolymer droplets that solidify when cured by an internal ultraviolet (UV) light source. It utilizes wax-like support material that allows the 3-D printer to print complex geometric features, such as curves and twists at the required resolution.

The main body of the phased-array antenna subsystem was printed in only two pieces; the top and bottom half parts were printed separately for ease of cleaning and metallization. A PolyJet resin, VeroWhite, was used for the main body. VeroWhite provides rigid 3-D prints that offer high resolution comparable to injection-molded polymer counterparts. The support material was mechanically removed and cleaned with a high-pressure water jet.

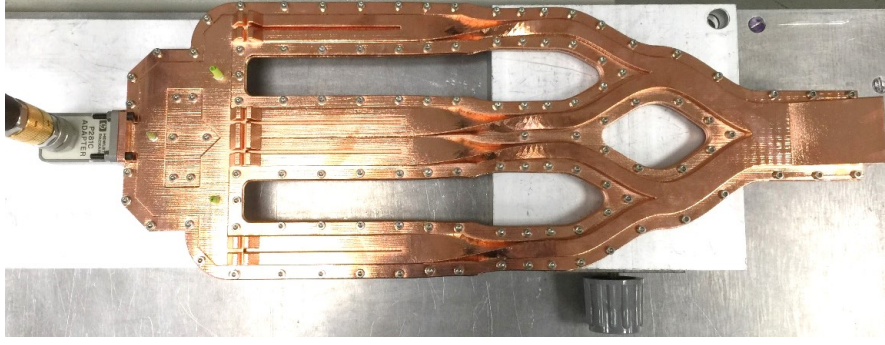
4.3.2 FDM

Fused deposition modeling 3-D printing was used to manufacture the dielectric inserts for the tunable phase shifters, their mounting brackets and calibrated screw mechanisms. The dielectric inserts were fabricated using an Ultimaker 2 printer, with Formfutura's Premium ABS (White) filament. Using a slicing software, Ultimaker Cura [8], the prints were set for a 100 μm layer height. The 3-D models of the inserts were orientated perpendicular to the build surface of the printer, in order to minimize the need for support structures. Infill parameters were customized to give a theoretical 100% infill density. The brackets and screw mechanisms were fabricated using an Original Prusa i3 MK3, with polylactic acid

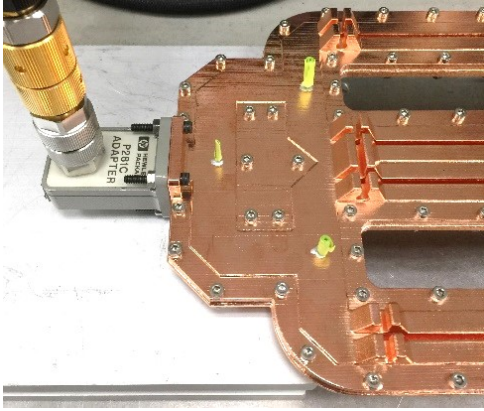
(PLA) filaments. The FDM parts were all lightly sanded down to increase their smoothness for ease in assembly and operation. Moreover, to reduce friction, some lubricant (petroleum jelly) was applied to the screws.

4.3.3 Metalizing

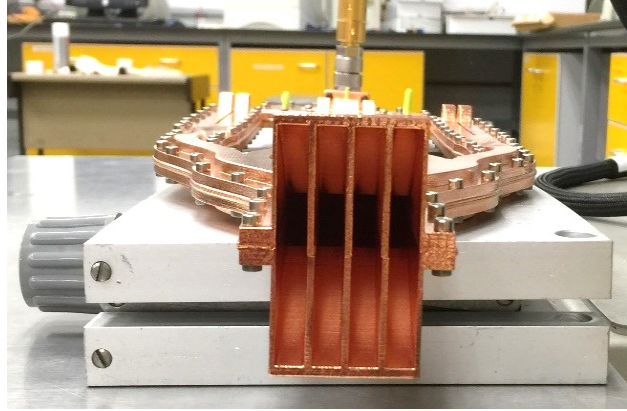
After cleaning, to metalize the main body of the phased-array antenna, a commercial copper-electroplating metallization process was employed. The electroplating process gave a copper layer thickness of 20 μm , corresponding to 38 skin depths (with $\delta = 0.53 \mu\text{m}$ at 15 GHz). The skin depth is given by $\delta = \sqrt{\rho_o / (\pi f_{min} \mu_o)}$, where the resistivity of copper $\rho_o = 1.72 \times 10^{-8} \Omega \cdot \text{m}$, f_{min} is the minimum operating frequency and μ_o is the permeability of free-space. The electroplated parts were assembled with standard M3 screws and nuts, as shown in Figure 4.8. Note that these M3 screws can be 3-D printed, or replaced altogether using a conductive adhesive, in order to achieve further weight savings.



(a)



(b)



(c)

Figure 4.8: Assembled phased-array antenna without dielectric insert components: (a) top view; (b) close-up view showing the waveguide port, power splitter pins, and phase shifter slot; and (c) close-up view showing aperture of horn antennas.

As weight reduction is one of the most appealing aspects of using polymer-based 3-D printing, the mass of the main body parts of the phased-array antenna was measured after the electroplating process. Predicted masses of conventionally machined solid-copper and aluminum phased-array antennas were obtained using CAD software (Autodesk Fusion 360) for comparison. The results are given in Table 4.1. As expected, polymer-based 3-D printing offers significant mass reduction when compared to conventional machining. Compared with conventional (i.e., non-3-D printing) commercial waveguide implementation, the mass reduction will be even greater, as multiple flanges are required to connect individual components together.

Table 4.1: Predicted main body mass comparison (without nuts and bolts) for 3-D printed and conventionally machined phased-array antenna

Manufacturing Method	Total Mass (g)	Conventional / 3-D Printed Mass Ratio
Conventional Aluminum	1026*	2.0
Conventional Copper	3397*	6.5
3-D Printed & Electroplated	522	-

* Predicted values from CAD software (Autodesk Fusion 360).

4.4 Measurements

Three sets of measurements were made at NPL: (i) input return loss; (ii) phase shift; and (iii) radiation pattern. The first two were taken using a vector network analyzer (Agilent PNA). Calibration standards with traceability to the International System of Units (SI), via their mechanical dimensions, were used to provide accurate S-parameter measurements with reference planes at the waveguide test ports. This was achieved by performing a Thru-Reflect-Line (TRL) calibration [9]. This calibration employed Ku-band waveguide standards: Thru – connecting the two VNA test ports together; Reflect – using a flush short-circuit connected, in turn, to each VNA test port; Line – using a $\frac{1}{4}$ -wave section of the waveguide. The length of the $\frac{1}{4}$ -wave line was 6.4 mm, which is the optimum length for calibration (i.e., by providing a phase shift of a $\frac{1}{4}$ of a guided wavelength) in the middle of the waveguide band (i.e., *ca.* 15 GHz for Ku-band waveguide).

The calibration was performed using an in-house calibration algorithm, employing a seven-term error-correction routine [10]. The complete measurement set-up (i.e., VNA, primary reference standards and calibration algorithm) is referred to as the NPL primary impedance microwave measurement system (PIMMS) [11], [12] and represents the UK's primary national standard system for S-parameter measurements.

The radiation patterns were measured in one of NPL's anechoic chambers, which comprises a transmit tower with reference antenna, a phi-over-theta spherical positioner system for the receiving antenna under test and associated RF test equipment operating from 0.5 to 220 GHz [13].

To accurately determine the gain of the antenna under test, it is critical to calibrate the gain of the reference antenna (standard gain horn) with low uncertainty. Prior to the radiation pattern measurement for the phased-array, the gain of the reference antenna (Narda 639B, having measured gains of 16.39 and 17.32 dBi at 15 and 17 GHz, respectively) was calibrated using the three-antenna extrapolation technique [14] with an uncertainty of 0.05 dB.

4.4.1 Return Loss Measurement

The input return loss for the complete phased-array antenna subsystem was measured as a function of dielectric insert penetration depth, across Ku-band, using the measurement set-up shown in Figure 4.9. Radar absorbing material (RAM) was placed in surrounding areas to minimize reflections. The measurement uncertainty for return loss was expected to be ± 0.2 dB at 20 dB.

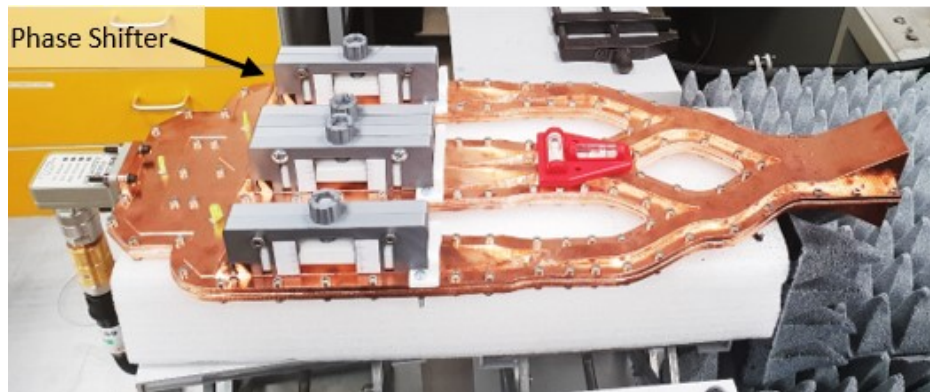


Figure 4.9: Return loss measurement set-up with dielectric inserts (fully assembled).

The EM simulated and measured return loss results across Ku-band are shown in Figure 4.10. The phased-array antenna shows excellent return loss behavior between 15 and 17 GHz. It can be seen that the measured return loss without any dielectric inserts is better than 12.9 dB across the 15 and 17 GHz operating

frequency range. When the dielectric inserts are at maximum depth, the worst-case return loss is 16.3 dB.

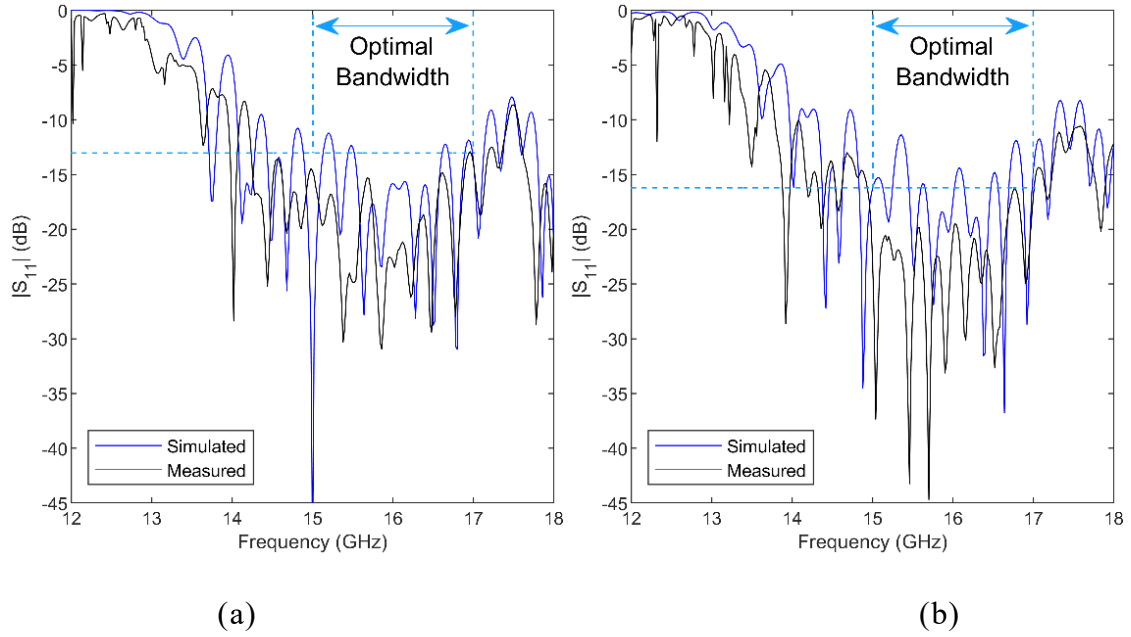


Figure 4.10: Simulated and Measured return loss of phased-array antenna across Ku-band: (a) without dielectric inserts; and (b) maximum dielectric insertion.

4.4.2 Phase Shifter Measurements

The relative phase shift of the phased-array antenna is the same as that for the individual phase shifters, when they are all in phase with one another. This was measured at 15 and 17 GHz, using the measurement set-up shown in Figure 4.11. The phased-array antenna and a receiving horn antenna were attached to fixed mounts with the boresight of the array aligned with the receiver horn. The separation distance between the two antenna apertures was 0.96 m (below the combined radiating far-field distance of 1.82 m). The measurement uncertainty for the transmission phase was expected to be 0.6° .

Using the screw mechanism, all four dielectric inserts were set to give the same level of penetration for each measurement. Starting from maximum insertion, measurements took place for every 0.5 mm of extraction. A Vernier caliper with the resolution of 0.01mm was used as a calibration reference. There was no significant drift observed between each measurement.

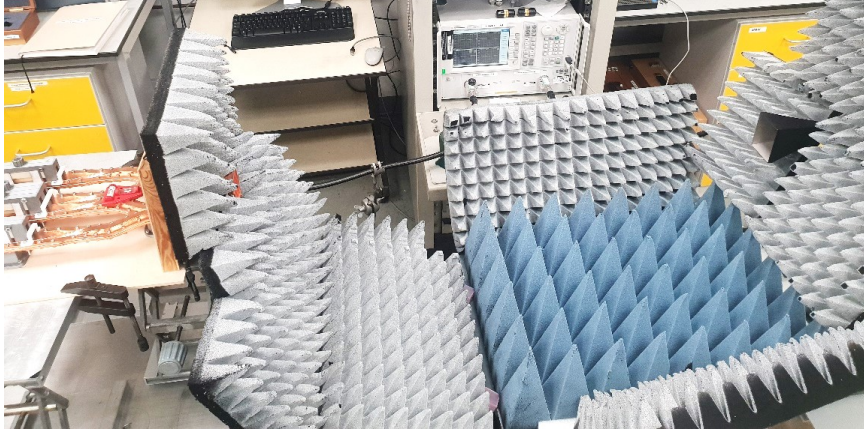


Figure 4.11: Relative phase shift measurement set-up.

The simulated and measured relative phase shift performance at 15 and 17 GHz is shown in Figure 4.12. Despite spatial uncertainty associated with measuring penetration depth and their relative alignments, the simulated and measured results show excellent agreement. From measured insertion phase $\angle S_{21}$, a maximum relative phase shift $\Delta\angle S_{21}(15 \text{ GHz}) = 263^\circ$ and $\Delta\angle S_{21}(17 \text{ GHz}) = 290^\circ$, which are within 0.4% and 0.7% of their respective predicted values.

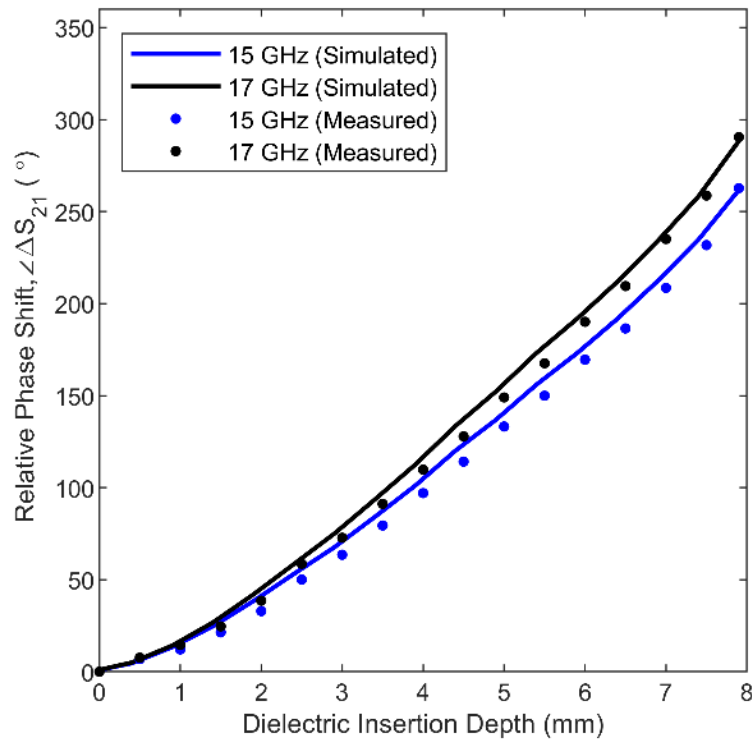
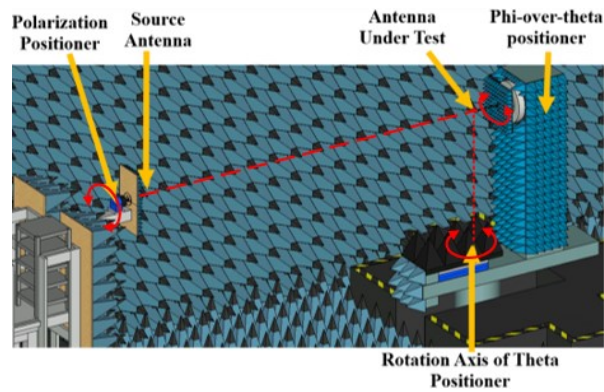


Figure 4.12: Simulated and measured relative phase shift at 15 and 17 GHz.

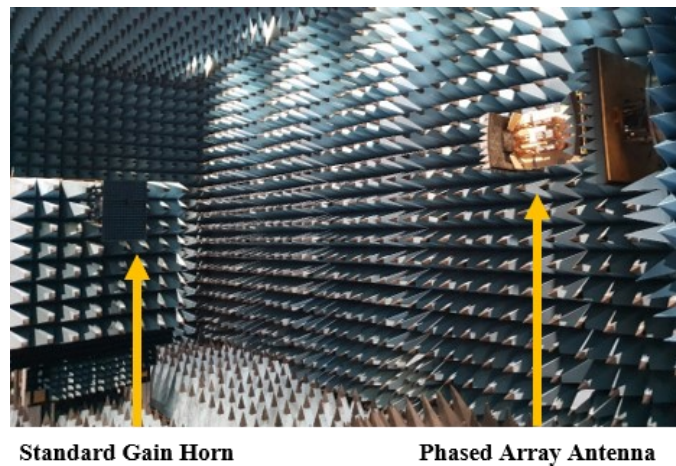
4.4.3 Radiation Pattern Measurements

The far-field E- and H-plane radiation patterns were measured at NPL's spherical test range. The spherical test range is housed within a temperature-controlled ($23 \pm 1^\circ\text{C}$), 15 m x 7.5 m x 7.5 m, fully screened anechoic chamber.

The phased-array antenna subsystem was mounted on a phi-over-theta spherical positioner, illustrated in Figure 4.13, with its aperture aligned over the rotation axis of the theta positioner. A WR-62 standard gain horn was used as the source antenna, which was installed on the polarization positioner. The separation distance between the source antenna and the phased-array antenna was 6 m.



(a)



(b)

Figure 4.13: NPL's anechoic chamber far-field antenna measurement set-up: (a) Standard measurement illustration; and (b) Photograph showing the phased-array antenna during H-plane measurements.

The combined radiating far-field (Fraunhofer) distance $R_{ff}(17 \text{ GHz}) = 2(D_{STD}^2 + D_{PAA}^2)/\lambda = 1.3 \text{ m}$, where $D_{STD} = 71 \text{ mm}$ and $D_{PAA} = 80 \text{ mm}$ are the maximum aperture size for the standard gain horn and phased-array antennas, respectively. The gain of the phased-array antenna was determined using a transfer method [15], by comparing it with that of the commercial standard gain horn antenna. The uncertainty for the antenna gain measurement was expected to be approximately $\pm 1 \text{ dB}$.

The penetration depths of the dielectric inserts were pre-calibrated to scan between the simulated target minimum and maximum beam steering angles $\theta_{max,min}$, in 5° increments. The phase difference between antenna elements $\Delta\phi(f)_{max}$, to achieve the maximum and minimum beam steering angles, was calculated using:

$$\Delta\phi(f)_{max} = \frac{\Delta\angle S_{21}(f)_{max}}{3} = \begin{cases} \pm 87^\circ & \text{at 15 GHz} \\ \pm 96^\circ & \text{at 17 GHz} \end{cases} \quad (4.10)$$

where, $\Delta\angle S_{21_{max}}$ is the simulated relative phase shift, at maximum dielectric penetration $g = 7.9 \text{ mm}$. The intermediate values of $\Delta\phi(f)$ for achieving the target beam steering angle θ were calculated using the true time steering expression in (4.2),

$$\Delta\phi(f) = \frac{d}{\lambda} \cdot \sin\{\theta\} \cdot 360^\circ \quad (4.11)$$

Using $\Delta\phi(f)$, the relative phase for the i^{th} phase shifter element was obtained from $\phi_i(f) = (i - 1) \cdot \Delta\phi(f)$, where $i \in \{1, 2, 3, 4\}$. The dielectric insert penetration depth corresponding to each value of $\phi_i(f)$ was interpolated from simulated relative phase shift given in Figure 4.12. From Figure 4.4(a), it can be seen that the phase shifter approximates to a true time shifter at 17 GHz, while having a transition characteristic at 15 GHz. The calibrated dielectric insert penetration depth settings are shown in Table 4.2.

Table 4.2 : Calibrated dielectric insert penetration depth settings

Target Beam Steering Angle, θ (°)	Phase Difference Between Antenna Elements, $\Delta\phi(f)$ (°)		Required Relative Phase Shift $\Delta\angle S_{21}$ (°) for Individual Phase Shifters								Dielectric Insert Penetration Depth, g (mm)			
	15 (GHz)	17 (GHz)	$\phi_1(f)$		$\phi_2(f)$		$\phi_3(f)$		$\phi_4(f)$		g_1	g_2	g_3	g_4
			15 (GHz)	17 (GHz)	15 (GHz)	17 (GHz)	15 (GHz)	17 (GHz)	15 (GHz)	17 (GHz)				
Minimum	-87	-96	262	288	175	192	87	96	0	0	7.9	5.9	3.5	0
-25	-76	-86	228	259	152	173	76	86	0	0	7.3	5.3	3.2	0
-20	-62	-70	185	209	123	140	62	70	0	0	6.2	4.5	2.7	0
-15	-47	-53	140	159	93	106	47	53	0	0	5.0	3.7	2.2	0
-10	-31	-35	94	106	63	70	31	35	0	0	3.7	2.7	1.7	0
-5	-16	-18	47	53	31	36	16	18	0	0	2.2	1.7	1	0
0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
5	16	18	0	0	16	18	31	36	47	53	0	1.0	1.7	2.2
10	31	35	0	0	31	35	63	70	94	106	0	1.7	2.7	3.7
15	47	53	0	0	47	53	93	106	140	159	0	2.2	3.7	5.0
20	62	70	0	0	62	70	123	140	185	209	0	2.7	4.5	6.2
25	76	86	0	0	76	86	152	173	228	259	0	3.2	5.3	7.3
Maximum	87	96	0	0	87	96	175	192	262	288	0	3.5	5.9	7.9

To ensure the dielectric inserts are in the right positions, small 3-D printed calibration blocks were placed at the bottom of the rails, on the phase shifter brackets, which prevent the dielectric inserts from penetrating further than the calibrated values during the measurements.

Measured far-field E-plane radiation patterns are shown in Figure 4.14. It is important to note that with the H-plane sectoral horn antennas, linearly polarized behavior is expected. During the measurement, the receiving standard gain rectangular horn antenna was set-up to have the same polarization direction as the phased array antenna. Therefore, the results are only showing co-polarization radiation patterns.

It can be seen that the total beam steering angle is 54° at 15 GHz and 52° at 17 GHz. By controlling the dielectric insert penetration depth, the phased-array antenna's beam angle can be adjusted to scan between -27° and $+27^\circ$ at 15 GHz and -25° and $+27^\circ$ at 17 GHz.

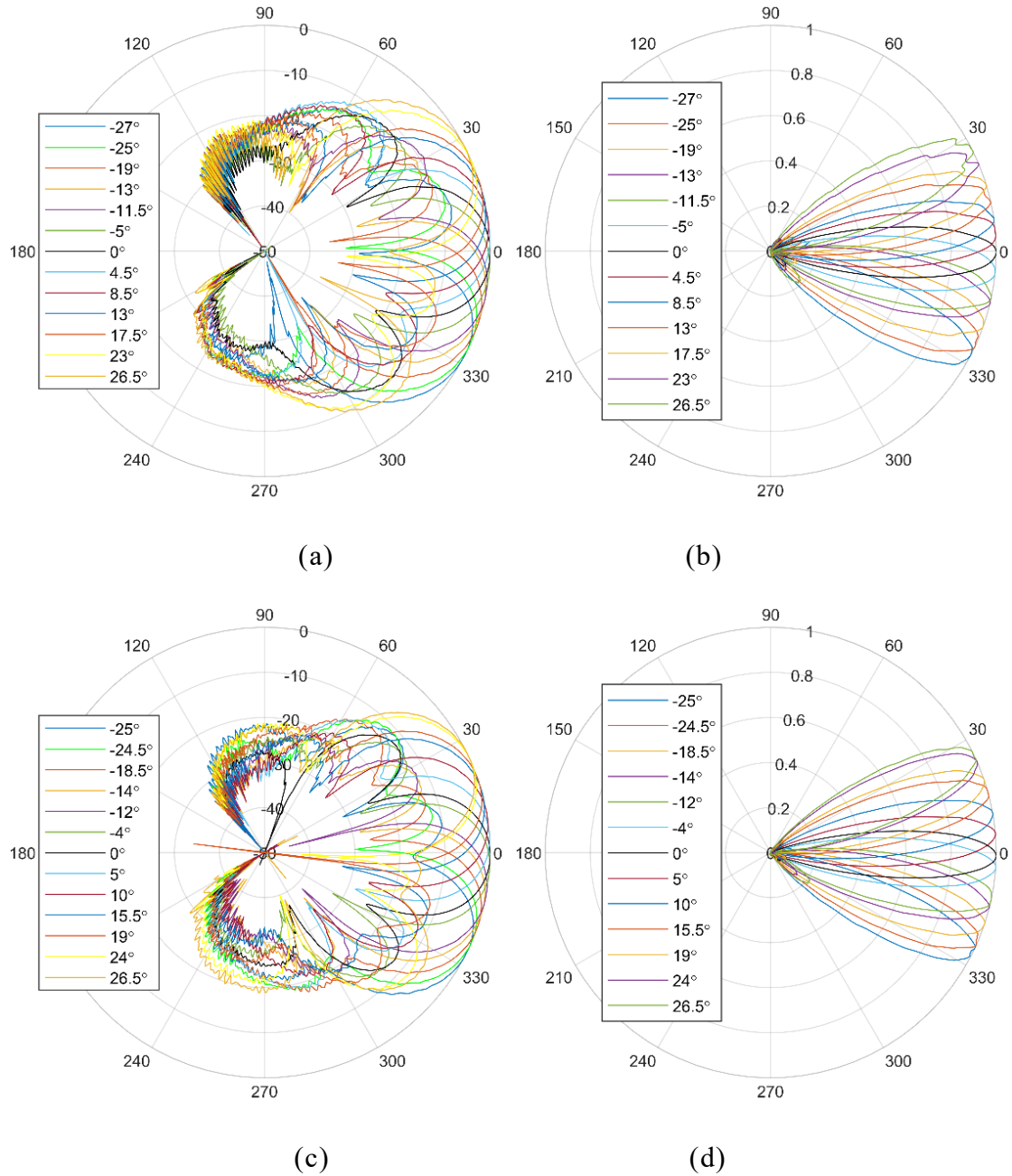


Figure 4.14: Measured far-field E-plane radiation patterns at 15 and 17 GHz, normalized to the broadside boresight value: (a) 15 GHz (logarithmic); (b) 15 GHz (linear); (c) 17 GHz (logarithmic); and (d) 17 GHz (linear).

Figure 4.15 shows predicted (theoretical and EM simulated) and measured beam steering angles from broadside boresight for each phase shift setting. The theoretical results can be obtained directly using either (4.1) or (4.2). For both 15 and 17 GHz, measured results show excellent agreement with the predicted results.

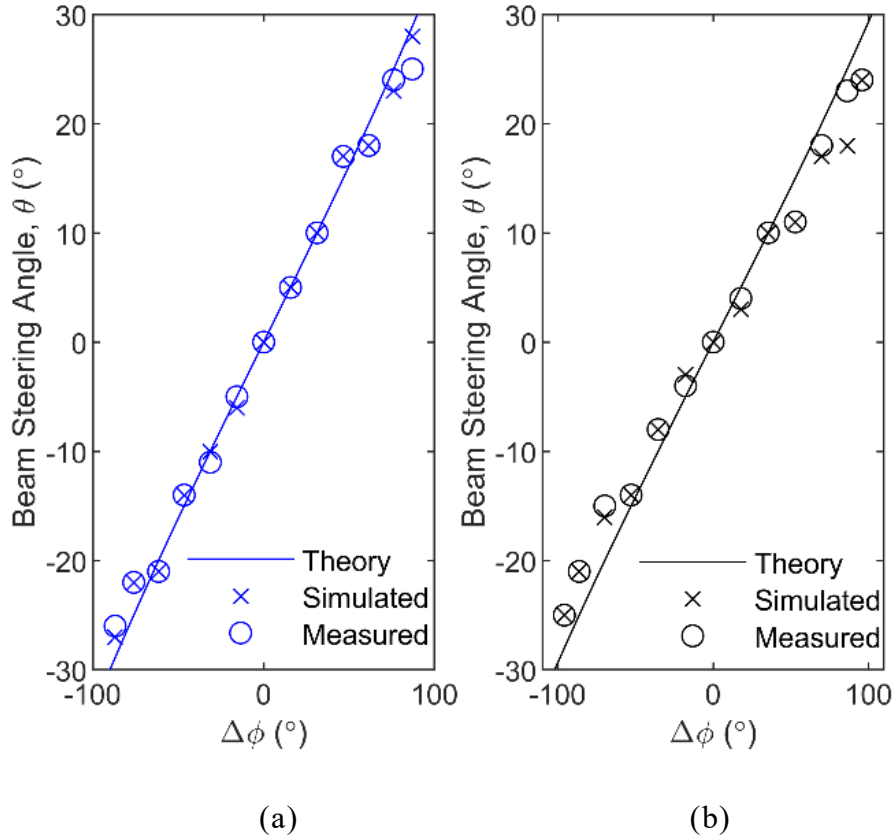
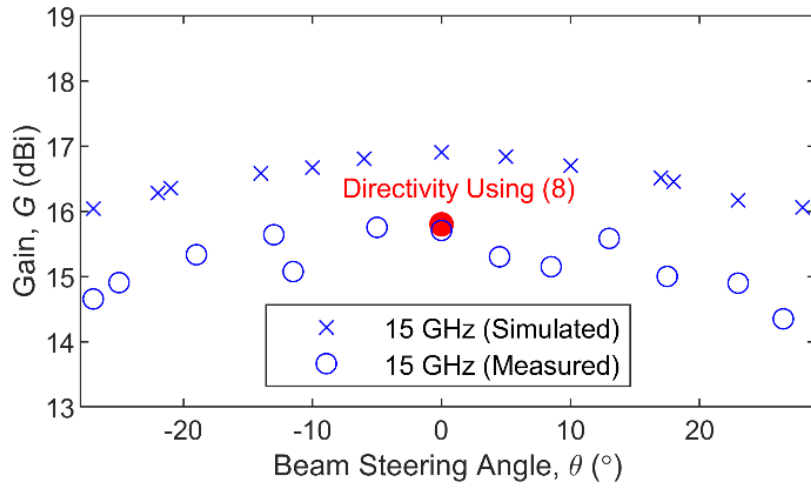


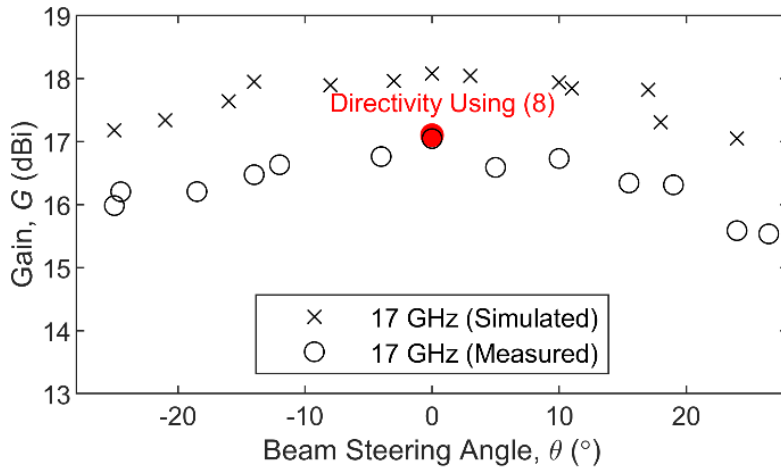
Figure 4.15: Theoretical, EM simulated and measured beam steering angles: (a) 15 GHz; and (b) 17 GHz.

The antenna gains for each target beam steering angle range from 14.4 to 15.8 dBi at 15 GHz and from 15.5 to 17.0 dBi at 17 GHz. At broadside boresight, the theoretical prediction and measured results show exceptional agreement; the measured gains of $G(15 \text{ GHz}) = 15.7 \text{ dBi}$ and $G(17 \text{ GHz}) = 17.0 \text{ dBi}$ are both only 0.1 dB lower than the calculated directivities. The EM simulated (with PEC metallization and no surface roughness or defects) and measured gains for the phased-array antenna are shown in Figure 4.16. There are approximately 1.5 dB and 1 dB differences at 15 GHz and 17 GHz, respectively. These discrepancies can be attributed to the finite conductivity of the metallization, surface roughness

and small defects [16]. To reduce the computational complexity of the whole phased array antenna model, perfect electric conductor (PEC) was used to model the outer shell of the antenna. As a result, the EM simulation does not consider conductor loss. This would be one of the major factors that caused the discrepancies between simulated and measured gain. Moreover, unlike PEC, electroplated copper with finite conductivity and imperfect metallization would cause some absorption within the main waveguide structure of the phased array antenna. This explains the reduced measured return loss values, as shown in Figure 4.10.



(a)



(b)

Figure 4.16: EM simulated and measured antenna gains at 15 and 17 GHz against beam steering angle: (a) 15 GHz; and (b) 17 GHz.

For each target value of $\Delta\phi(f)$, given in Table 4.2, the EM simulated and measured E-plane field intensity are shown in Figure 4.17. The E-field intensity peak represents the location and size of the main lobe and any sidelobes. It can be seen that for both 15 GHz and 17 GHz, simulated and measured results have similar beam profiles, with approximately the same sizes and locations of the main lobe and sidelobes.

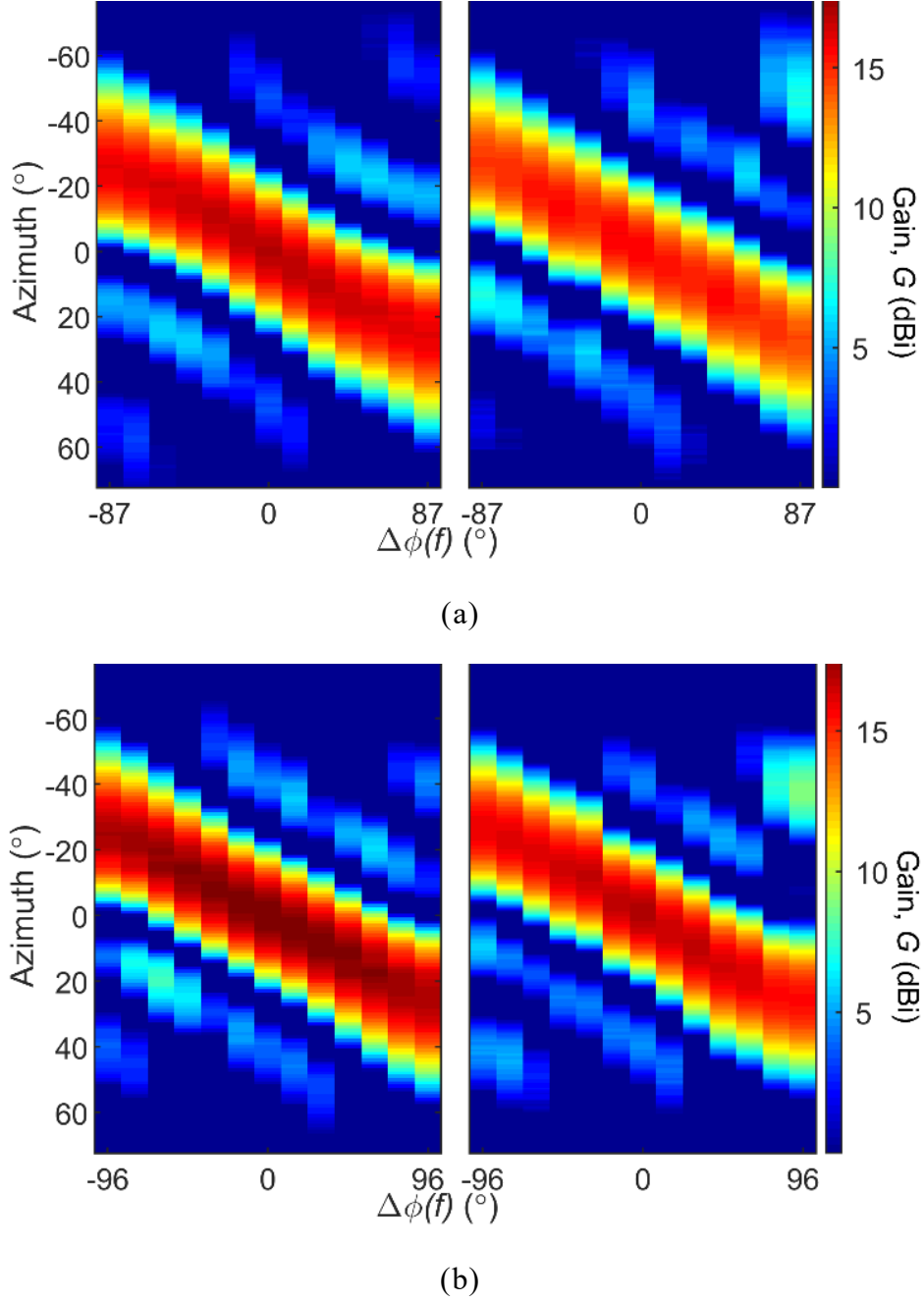
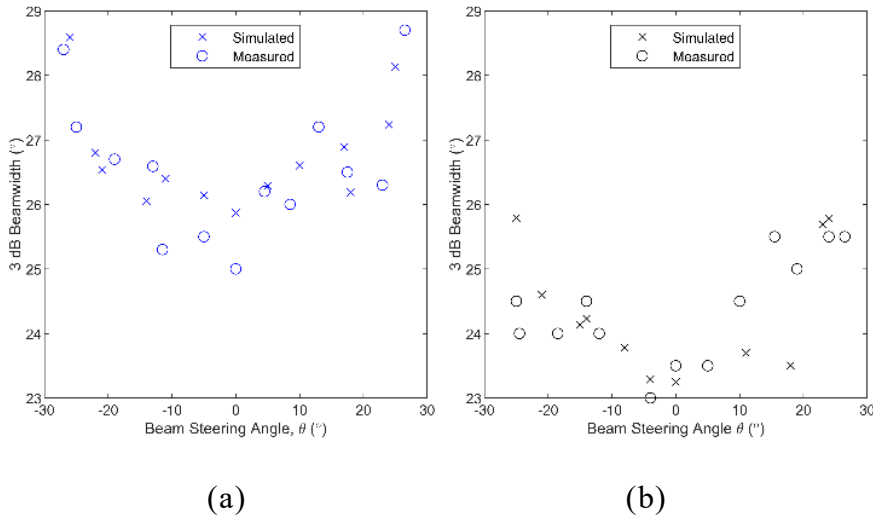


Figure 4.17: EM simulated (left) and measured (right) radiated E-plane field intensities: (a) 15 GHz; and (b) 17 GHz.

The EM simulated and measured E-plane radiation pattern properties are shown in Figure 4.18. The 3 dB beamwidths for each beam steering angle are shown in Figure 4.18(a) and 18(b). The phased-array antenna shows almost identical performance to predictions at 15 GHz and 17 GHz. The maximum measured beamwidths are 29° and 26° at 15 GHz and 17 GHz, respectively. The minimum measured beamwidths are 25° and 23° at 15 GHz and 17 GHz, respectively.

The worst-case sidelobe levels for each beam steering angle are shown on Figure 4.18(c) and 18(d). There is close agreement between the simulated and measured sidelobe levels. At 15 GHz, more than half of the measured sidelobe levels are below -10 dB, but higher than the simulated results. At 17 GHz, the results improve with most of the measured sidelobe levels below -10 dB. The worst-case measured sidelobe levels are -7.3 dB and -6.7 dB at 15 GHz and 17 GHz, respectively. The 3 dB beamwidth and sidelobe levels can be reduced by further optimizing the dielectric penetration depth and improving the control accuracy of the dielectric insertion mechanism.



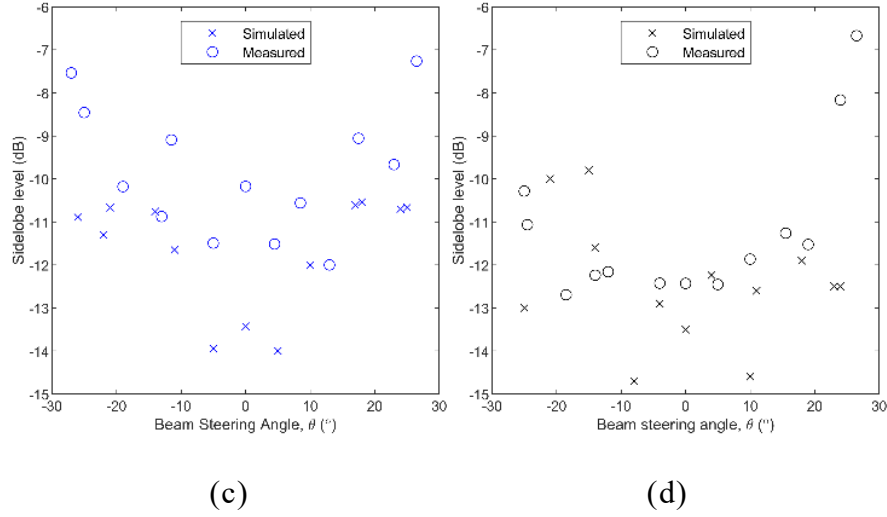


Figure 4.18: EM simulated and measured E-plane radiation pattern properties: (a) 3 dB beamwidths at 15 GHz; (b) 3 dB beamwidths at 17 GHz; (c) worst-case sidelobe levels at 15 GHz; and (d) worst-case sidelobe levels at 17 GHz.

Finally, far-field H-plane radiation patterns were measured at broadside boresight without the dielectric inserts. The EM simulated and measured results are shown in Figure 4.19. The measured 3 dB beamwidths are 24° and 22° at 15 GHz and 17 GHz, respectively, which are nearly identical to the simulated results. It is interesting to note that the 3 dB beamwidths for the complete phased-array antenna, at broadside boresight, for the H-plane are approximately only 1° less than those for the E-plane.

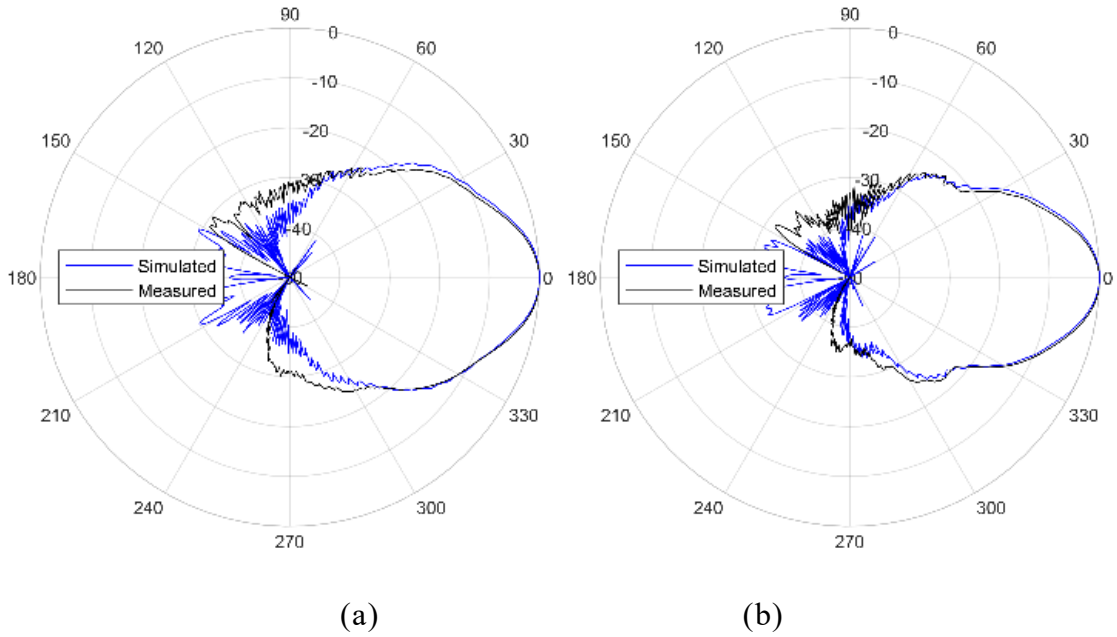


Figure 4.19: EM simulated and measured far-field H-plane radiation patterns: (a) 15 GHz (logarithmic); and (b) 17 GHz (logarithmic).

These results show that even with finite conductivity walls, rough surfaces and defects, the manufactured 3-D printed phased-array's performance is comparable to that of an ideal model having PEC metallization and no surface roughness or defects.

4.5 Conclusion

This work has introduced the first fully 3-D printed tunable microwave subsystem. The Ku-band 4-element phased-array antenna was manufactured using polymer-based 3-D printing technologies, consisting of 26 circuit elements, and demonstrated excellent measured performance. A combination of industrial-level PolyJet technology and entry-level FDM was used to print the phased-array antenna. By independently adjusting the depths of the dielectric inserts, the beam was fully steerable from -27° to $+27^\circ$ at 15 GHz and -25° to $+27^\circ$ at 17 GHz. Measured beam profiles and antenna gains show excellent agreement with simulated results.

4.6 Acknowledgements

Work presented in this chapter was published on the paper: "Polymer-based 3-D printed Ku-band steerable phased-array antenna subsystem," by **S.-H. Shin**, D. Alyasiri, M. D'Auria, W. J. Otter, C. W. Myant, D. Stokes, Z. Tian, N. M. Ridler, and S. Lucyszyn (*IEEE Access*, vol. 7, pp. 106662-106673, Aug. 2019) [1]. The research on this work has been performed under the supervision of Prof. S. Lucyszyn and he went over every word of the original article. D. Alyasiri, M. D'Auria, W. J. Otter started the project with the initial design. PolyJet fabrication was undertaken within the Department of Design Engineering at Imperial College by C. W. Myant. Measurements at NPL were performed under the supervision and help of D. Stokes, Z. Tian, and N. M. Ridler.

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5 3-D Printed G-band Quasi-Optical Components

5.1 Introduction

This chapter presents the characterization of 3-D printed G-band quasi-optical components and subsystems based on a published article by Shin *et al.* [1]. Here, five different quasi-optical component types are investigated; rectangular horn antennas, 90° off-axis parabolic mirrors, radiation absorbent material (RAM), grid polarizers and dielectric lenses. Performance of 3-D printed antennas, mirrors and RAM are compared against their commercial equivalents. Also, a 3-D printed housing with an axial rotation platform is also designed and fabricated for performing two-port, quasi-optical, horn-mirror-polarizer-mirror-horn subsystem measurements. The ‘plug and play’ housing give a fast measurement set-up, while minimizing misaligning losses. By analyzing each component and the fully assembled subsystem, it opens up new opportunities for low-cost fabrication of quasi-optical upper-millimeter-wave subsystems and components.

5.2 Design

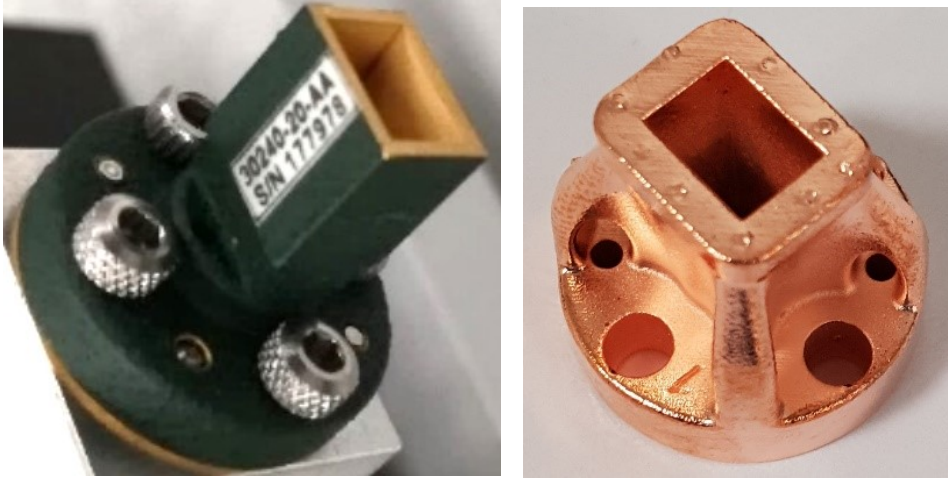
5.2.1 Horn Antennas

A commercial Flann Microwave Ltd 20 dBi standard gain rectangular horn antenna (model #30240-20) is chosen [2] for replication. This is commercially manufactured by copper electroforming on precision mandrels, to ensure high accuracy and repeatability. These standard gain antennas are very well understood and there is a comprehensive datasheet available enabling replica 3-D printed versions to be made [2].

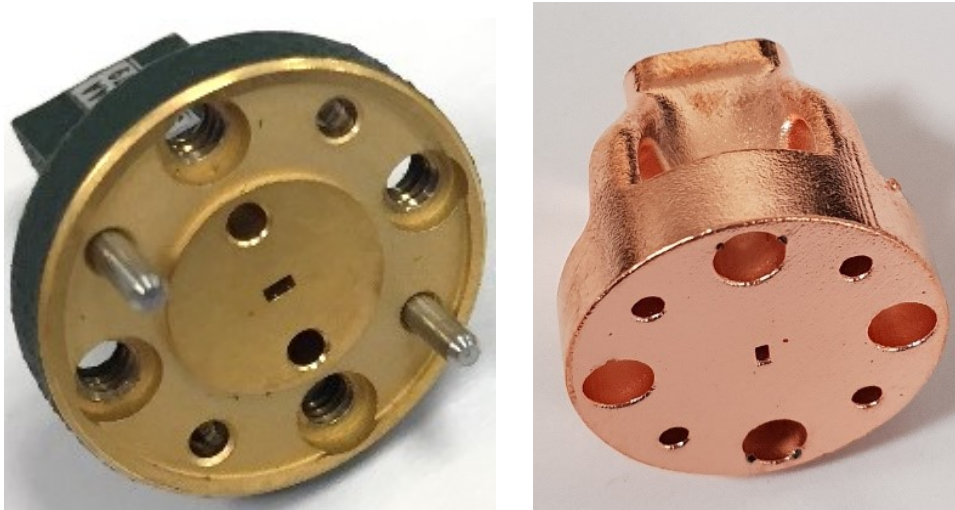
For G-band, the rectangular waveguide has WR-5 standard internal aperture dimensions of only $1,295 \mu\text{m} \times 647.5 \mu\text{m}$ (this is equivalent to the WM-1295 waveguide size in the IEEE standard). The linearly tapered horn antenna has a flare length of 14.5 mm, with maximum internal aperture dimensions of 6.8 mm $\times$ 5.0 mm, and an overall length of 17.5 mm. From these dimensions, assuming infinitesimally thin metal walls and a perfect electrical conductor (PEC), the theoretical directivity is $4\pi A_{max}/\lambda_0^2$ for any aperture antenna, giving 20.85 dBi

at 160 GHz, where the maximum internal aperture area $A_{max} = 34 \text{ mm}^2$ and free-space wavelength $\lambda_0(160 \text{ GHz}) = 1.8737 \text{ mm}$. As previously observed in Chapter 4, in practice, the reduction in gain due to finite wall conductivity can be effectively compensated for by the reduction in lost power in the backlobes (since backward radiation leakage caused by knife-edge diffraction is reduced with finite wall thickness). At 160 GHz, the calculated far-field (or Fraunhofer/Rayleigh) distance $z_{ff} = 2D_{HA}^2/\lambda_0(160 \text{ GHz}) = 76.0 \text{ mm}$, with the largest internal aperture dimension of the horn antenna $D_{HA} = 8.44 \text{ mm}$.

To increase structural rigidity and reduce structural stress concentrations at sharp corners, regions with relatively thin walls are thickened and sharp corners rounded. In addition, buttresses have been added between the horn and flange, to reduce warpage caused by the post-UV curing process. The flange design is based on IEEE standard specifications [3]. Figure 5.1 shows the fabricated ruggedized 3-D printed modified replica horn antenna.



(a)



(b)

Figure 5.1: Photographs of commercial counterpart (left) and ruggedized 3-D printed modified replica (right) G-band rectangular horn antennas: (a) showing horn aperture; (b) showing flange aperture.

5.2.2 90° Off-axis Parabolic Mirrors

A commercial Edmund Optics Ltd 90° OAPM (original model #63-188, with #35-522 replacement) is chosen for benchmarking against, with specifications given in Table 5.1 [4] and the associated ray tracing illustrated in Figure 5.2.

Table 5.1: Commercial 90° off-axis parabolic mirror specifications [4].

Parameters	Values
Mirror Diameter, D	76.2 mm
Off-Axis Angle	90°
Parent Focal Length, f_{pa}	38.1 mm
Y Offset, y_{off}	76.2 mm
Reflected Effective Focal Length, f_{eff}	76.2 mm
Substrate and Coating Material	Protected Aluminum
Intended Optical Wavelength Range	400 to 700 nm
RMS Surface Roughness	17.5 nm



(a)



(b)



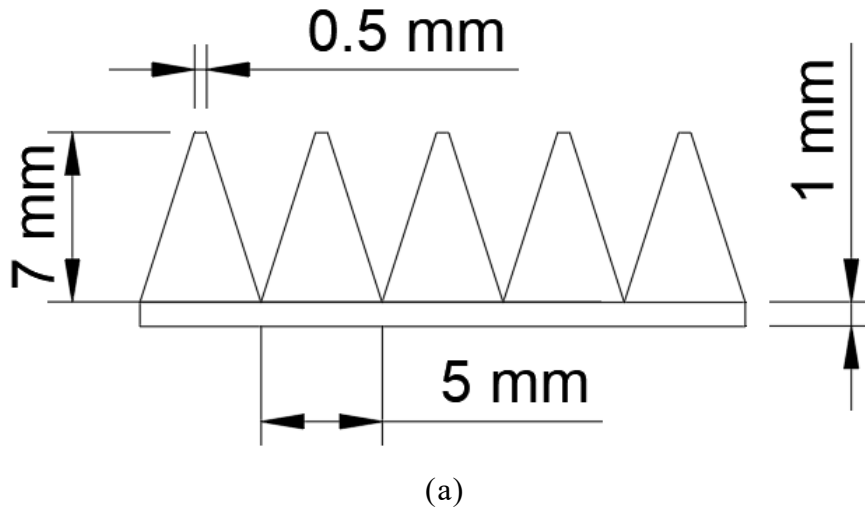
(c)

Figure 5.3: Photographs of commercial counterpart (left) and mass-reduced 3-D printed modified replica (right) 90° off-axis parabolic mirrors: (a) isometric views; (b) front views; and (c) side views.

5.2.3 Radiation Absorbent Material

Ultra-wideband RAM can be realized using naturally absorbing materials and wave impedance transforming geometries. One common example is carbon-loaded polymer shaped into a 2-D array of pyramids. Here, tessellating pyramidal-array RAM is FDM 3-D printed using Proto-pasta conductive PLA. This polymer composite is made with Natureworks 4043D PLA, a dispersant and 20% conductive carbon black powder [5].

Due to the nature of FDM 3-D printing, the tips of the pyramids cannot be infinitesimally small. In practice, the top of the pyramid has dimensions of $0.5 \text{ mm} \times 0.5 \text{ mm}$. The base of the pyramid has dimensions of $5 \text{ mm} \times 5 \text{ mm}$, which is large enough to be structurally stable during 3-D printing. The height of the pyramid is 7 mm , which is long enough to provide a gentle taper into the absorbing medium, enabling operation down to microwave frequencies. Finally, a 1 mm thick base is added to provide increased absorption (and reduced transmission) and structural stability. Figure 5.4 shows the key dimensions and a fabricated tile for the FDM 3-D printed RAM.





(b)

Figure 5.4: FDM 3-D printed pyramidal RAM: (a) cross-sectional CAD drawing; and (b) photograph.

5.2.4 Grid Polarizer

Ideally, the strip width, thickness and separation distance needs to be as small as possible. With the linear polarizer, the $400\text{ }\mu\text{m}$ strip width is defined by the FDM 3-D printer's nominal minimum extrusion line width of $400\text{ }\mu\text{m}$. To avoid strips touching, while allowing variations in extrusion line widths and gold-leaf gilding, the chosen strip separation distance is $h = 800\text{ }\mu\text{m}$. As a result, the theoretical TE upper cut-off frequency is $f_c = c/2h = 187.37\text{ GHz}$, where c is the speed of light in free space.

For structural rigidity, eight extrusion lines are printed on top of one another, giving a total thickness of $800\text{ }\mu\text{m}$. On top of this, three layers of 24 carat (fineness of 99.99) gold-leaf is gilded, giving a total metallization thickness of

approximately $0.36 \mu\text{m}$. At the worst-case G-band frequency of 140 GHz, the skin depth $\delta = 1/\sqrt{\pi f \mu_o \sigma_o} = 0.21 \mu\text{m}$, where μ_o is the permeability of free space and the bulk conductivity of gold at DC is $\sigma_o = 4.1 \times 10^7 \text{ S/m}$. Therefore, with a total metallization thickness of only $T_m = 1.714\delta$ (instead of the recommended minimum thickness of 5δ), limited grid polarizer performance is to be expected from the ultra-low cost implementation. For example, as a flat continuous gold sheet, the absorption (or penetration) loss of electromagnetic energy as it propagates through the three sheets is $-10 \log_{10}\{e^{-2T_m/\delta}\} = 15 \text{ dB}$; while its shielding effectiveness would be much higher. Figure 5.5 shows the design and final realization of the grid polarizer.

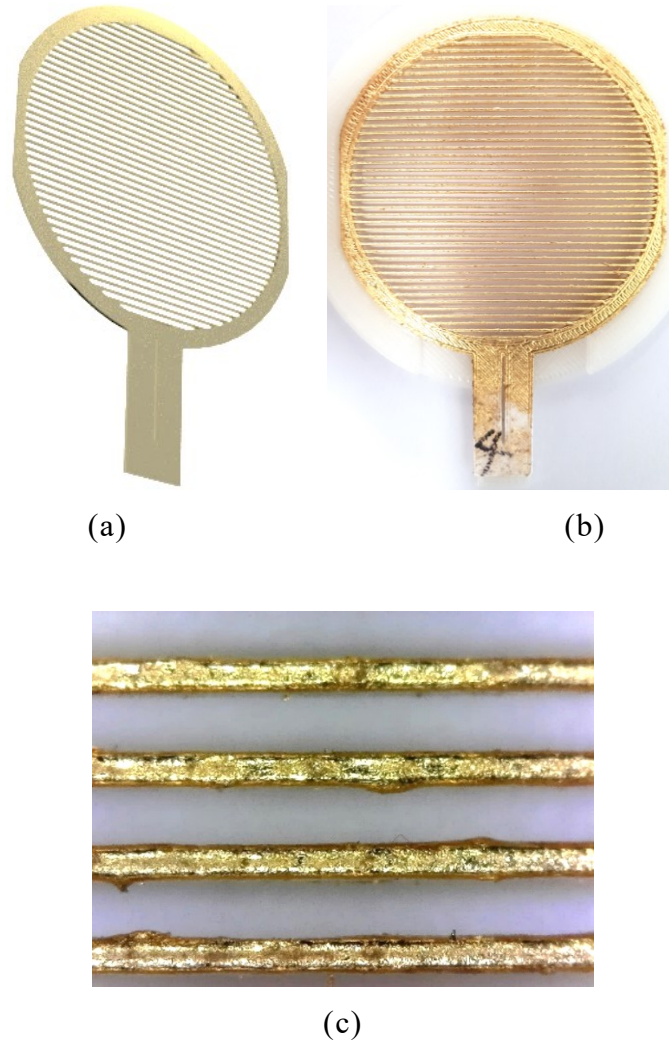


Figure 5.5: 3-D printed and gold-leaf gilded grid polarizer: (a) CAD rendering; (b) photograph; and (c) zoomed-in view photograph.

5.2.5 Quasi-optical Housing

The fully 3-D printed quasi-optical housing is designed for two-port measurements, with a horn-mirror-mirror-horn (HMMH) path and a two-axis rotational platform at the center to support the devices under test (DUT). With reference to Figure 5.6, the entire housing frame is FDM 3-D printed as a single piece, having maximum outer dimensions of 270 mm $\times$ 219 mm $\times$ 102 mm. Two rectangular slots are provided for the horn antennas, and their mechanical interface adapters, associated with Ports 1 and 2 of the vector network analyzer (VNA) and its frequency extension heads.

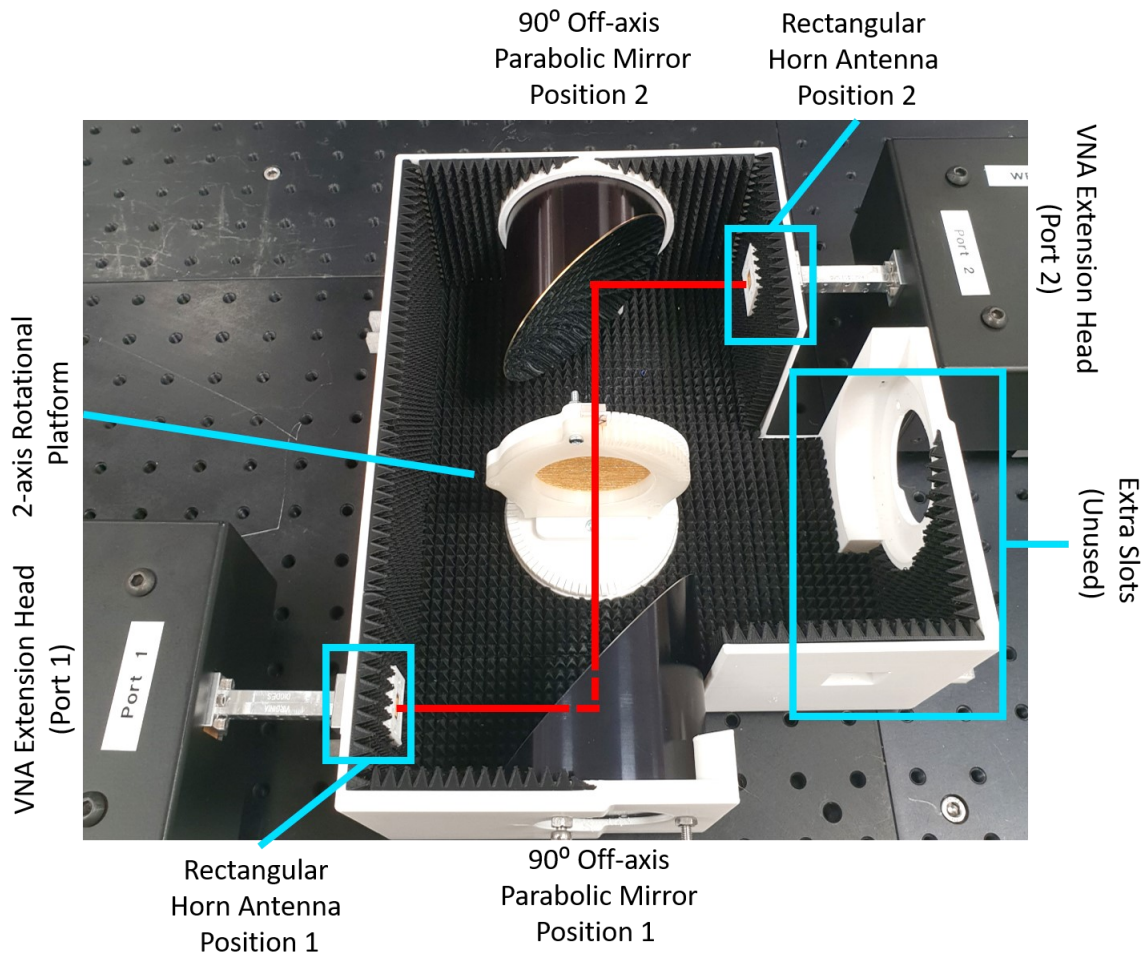


Figure 5.6: 3-D printed quasi-optical housing with commercial horns and mirrors assembled here for polarizer measurements (signal path indicated in red).

Circular slots are also provided for up to three 90° off-axis parabolic mirrors. In Figure 5.6, two mirrors are shown, which collimate the diverging incident beams from the horn antennas, with the DUT's rotating platform between mirrors. This platform is intended for measuring the polarizer, as shown in Figure 5.6, and RAM. As mentioned previously, the inner walls of the housing are lined with the FDM 3-D printed RAM.

The spatial design of the housing was the result of a number of practical constraints and required the use of geometrical ray tracing and full-wave electromagnetic (EM) simulations.

For example, the orientation of the HMMH arrangement is important. It is recommended that the two OAPMs have coinciding axes for the parent parabolas [6]; this optimal arrangement results in minimal aberrations due to source misalignments. As a result, the frequency extension heads need to be located side-by-side, such that both reference ports are facing the same direction. However, a VNA and its frequency extension heads are calibrated such that both reference ports are in-line and facing each other in opposite directions.

To reduce instrumentation errors associated with the coaxial cables (local oscillator, RF and reference signals) connecting the VNA to the frequency extension heads, Virginia Diodes (VDI) recommends that cable bending and movement should be minimized to avoid introducing significant phase mismatching; with sensitivity increasing as wavelength decreases. For this reason, the orientation of the HMMH arrangement is such that the two OAPMs are rotated 180° about their common optical axis [6]; being more susceptible to aberrations due to source misalignments. Therefore, housing slots for the rectangular horn antennas are orientated to face in opposite directions, as shown in Figure 5.6.

Another consideration is the spatial distance between mirrors since the distance between the horn and associated mirror is fixed by the focal length. Since there is a DUT placed between mirrors, which may be rotated, there has to be sufficient separation distance. Conversely, the maximum distance of 207.6

mm, between center points, is dictated by the maximum build volume capacity of the FDM 3-D printer.

To verify the focal positions of the HMMH arrangement, 2-D ray tracing is undertaken with MATLAB coding, and the results are shown in Figure 5.7(a). Clearly, with the arrangement, assuming an ideal point-source and perfect alignment, the beam diverges, collimates and converges perfectly with identical focal distances of 76.2 mm at each port.

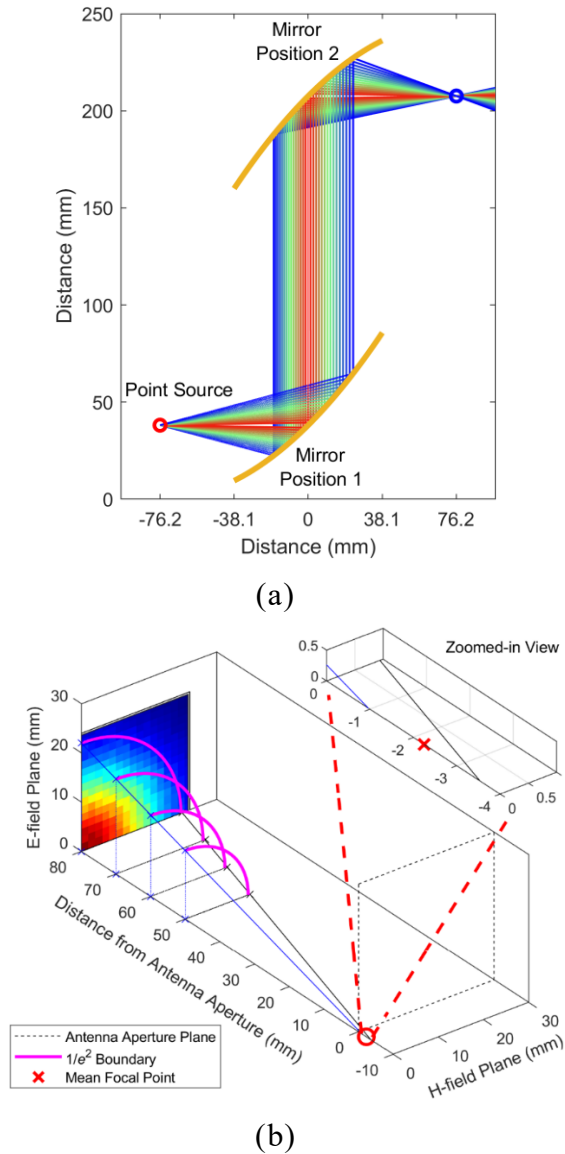


Figure 5.7: Ray tracing for determining quasi-optical component positions: (a) the HMMH arrangement having the two OAPMs rotated 180° about their common optical axis; and (b) effective equivalent focal point inside the rectangular horn antenna.

In practice, with quasi-optics, the minimum insertion loss is obtained when the positioning of all the HMMH components are optimized. A horn antenna does not represent the ideal point source. For this reason, the effective equivalent point source and focal point locations within the flare of the horn antenna must be established. To do this, EM simulations using HFSS and ray tracing are combined. With the former, a 2-D slice of the radiating horn antenna boresight beam profile is recorded at a distance of 80 mm from the radiating aperture. Then the $1/e^2$ beamwidth boundaries are determined for both the E- and H-field planes. To speed up this process, only 1-D slices are recorded in the E- and H-field planes, at distances of 50, 60 and 70 mm. Finally, rays are traced back to the propagation z-axis. Note that the $1/e^2$ beamwidths against z-axis distance get closer to a linear relationship as z increases; with values from 50 to 80 mm being large enough to undertake this extrapolation. As shown in Figure 5.7(b), the results reveal optimal focal points at 0.97 mm and 3.5 mm inside the horn antenna for the E- and H-field planes, respectively. For this reason, the slots for the horn antennas are positioned to be aligned near the arithmetic mean value of 2.25 mm, between these two focal points.

5.2.6 Dielectric Lenses

To design dielectric lenses, the material first needs to be characterized. Elegoo Water Washable Rapid Resin (Ceramic Grey) is the material used here, as it is capable of producing parts to relatively very high accuracy with ultra-low cost 3-D printing. To characterize the dielectric properties at G-band, 4 mm and 6 mm thick disk samples are MSLA 3-D printed and tested with the SWISSto12 WR-5 material characterization kit (MCK) at the UK's National Physical Laboratory (NPL) [7]. The extracted dielectric constant ϵ_r and loss tangent across G-band is shown in Figure 5.8.

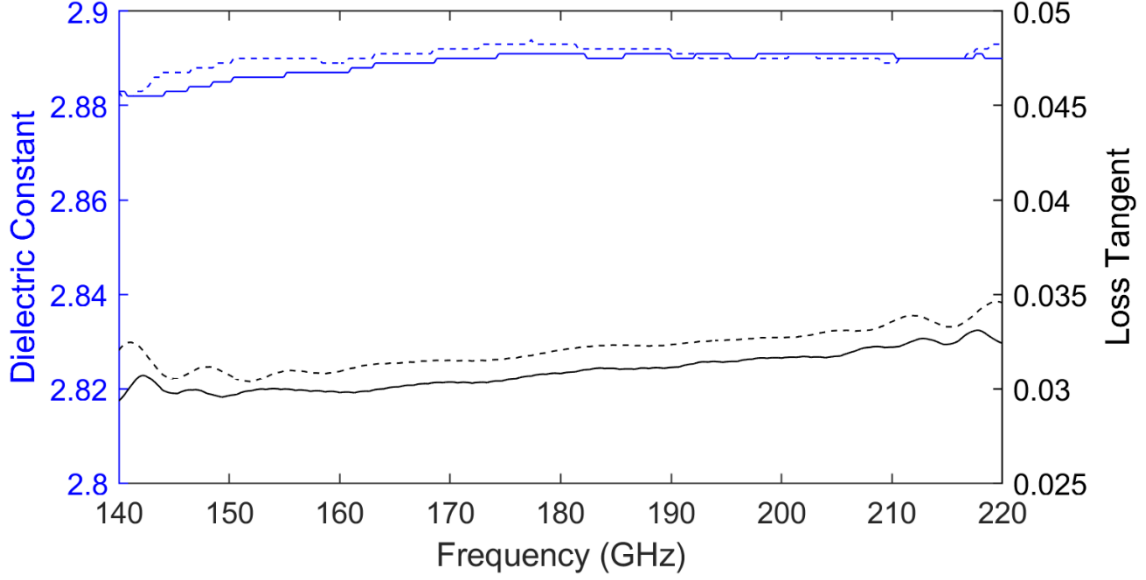


Figure 5.8: Extracted dielectric constant (blue lines) and loss tangent (black lines) for the 4 mm (dotted lines) and 6 mm (solid lines) samples.

Using the G-band average dielectric constant value of $\epsilon_r = 2.89$ (with refractive index $n \simeq \sqrt{\epsilon_r} = 1.7$), a hyperbolic aspherical plano-convex lens is designed. In theory, this type of lens can provide a perfect focal point without spherical aberration. The curvature profile of the lens can be derived using textbook expressions. For any given lens radius r , the free-space optical path length from the curvature of the lens to the focal point $OPL(r)$ is given by:

$$OPL(r) = \sqrt{\left(f_{pl} + z(r)\right)^2 + r^2} \quad (5.1)$$

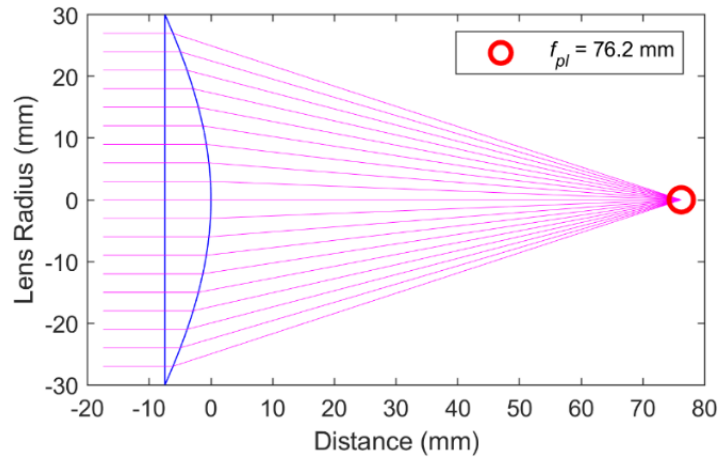
where, the target focal path length f_{pl} is the minimum distance from the lens to the focal point (along the optical axis) and $z_{max} - z(r)$ represents the lens curvature profile with z_{max} representing the maximum thickness of the lens (with $r = 0$). The optical path length at the center of the lens (OPL_c) is:

$$OPL_c = f_{pl} + \sqrt{\epsilon_r} \cdot z(r) \quad (5.2)$$

Hence, equating (5.1) and (5.2) gives:

$$z^2(r)(\epsilon_r - 1) + 2f_{pl}z(r)(\sqrt{\epsilon_r} - 1) - r^2 = 0 \quad (5.3)$$

Using the quadratic expression in (5.3), $z(r)$ can be extracted as a function of radius, up to the chosen maximum target value of $r = r_{max}$. Using Snell's law, a MATLAB code is written for a target focal path length $f_{pl} = 76.2$ mm and a lens diameter of $2r_{max} = 60$ mm, giving $z_{max} = 7.45$ mm, with optical ray tracing results shown in Figure 5.9(a). This clearly shows that the profile provides a perfect focal point without any spherical aberration. In practice, to minimize warpage during the post-UV curing process, a 1 mm thick base and a 2 mm thick rim at the edge of the lens are added. The key spatial dimensions and a photograph of the 3-D printed hyperbolic aspherical plano-convex lens are shown in Figure 5.9(b) and (c), respectively.



(a)



(b)



(c)

Figure 5.9: Hyperbolic aspherical plano-convex conventional lens: (a) optical ray tracing diagram; (b) cross-sectional CAD drawing; and (c) photograph.

Using the same hyperbolic profile described previously, a plano-convex Fresnel lens is designed. The thickness of the original lens is sliced into seven 2 mm thick layers, which are then collapsed to create the characteristic Fresnel lens cross-sectional profile. Since the basic Fresnel lens thickness does not exceed 2 mm, the upper limit on its diameter is not dictated by dielectric losses, as would be the case with conventional lenses. For this reason, the Fresnel lens diameter is increased to 80 mm. In practice, a 1 mm thick base and a 2.4 mm thick rim at the edge of the lens are added to minimize warpage. The key spatial dimensions and a photograph of the 3-D printed Fresnel lens are shown in Figure 5.10(a) and (b), respectively.

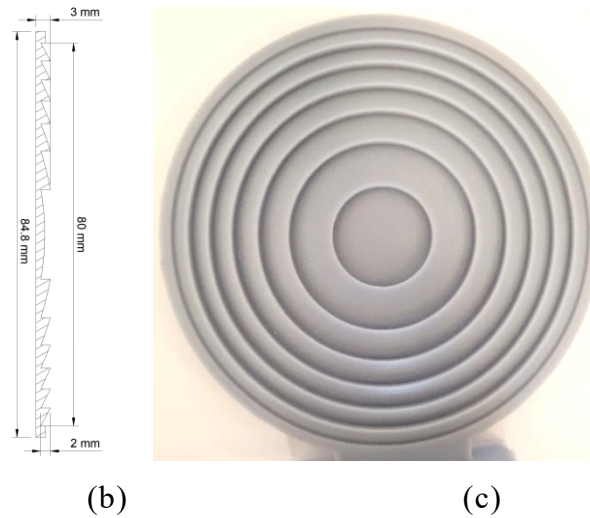


Figure 5.10: Hyperbolic aspherical plano-convex Fresnel lens: (a) cross-sectional CAD drawing; and (b) photograph.

5.3 Fabrication

Two different generic types of 3-D printing technologies are used in this work: (i) vat photopolymerization with UV cured photopolymer resin, employing both MSLA and SLA; and (ii) FDM with PLA-based thermoplastic filaments. All the 3-D printed CAD models are drawn using Autodesk Fusion 360, with a maximum triangular mesh refinement setting to maintain the integrity of curved surfaces.

5.3.1 Vat Photopolymerization

To fabricate the rectangular horn antennas, a low-cost desktop MSLA printer, Elegoo Mars, is employed with upgraded linear rails and bearings (costing approximately \$350). This printer utilizes a 2K resolution UV liquid crystal display (LCD) screen, as a mask for illuminating and curing the resin. This provides prints with ultra-high resolution (quoted by the manufacturer as $47\text{ }\mu\text{m} \times 47\text{ }\mu\text{m} \times 20\text{ }\mu\text{m}$ [8]) and well-defined corner and edge features. However, due to its relatively small build volume (with a maximum dimension of $120\text{ mm} \times 68\text{ mm} \times 155\text{ mm}$ [8]) and visible voxel boundaries (with certain print scenarios), it is not optimal for the 90° OAPM.

3-D CAD models for the rectangular horn antenna and the lenses are set to a $20\text{ }\mu\text{m}$ layer-height using the Chitubox slicer software [9]. The slicer also adds support material to some outer surfaces of the rectangular horn antenna. Internal and flange surfaces are avoided, to ensure maximum dimensional accuracy and to minimize the need for mechanical post-processing. Elegoo Water Washable Rapid Resin (Ceramic Grey) is used, as it provides high precision and low shrinkage.

After the 3-D printing has completed, the parts are post-processed at room temperature (to minimize warpage) with the following steps: (i) pre-rinse in a water bath at room-temperature; (ii) mechanical support structure removal; (iii) water-jet cleaning; (iv) 30 minutes in an ultrasonic bath, to remove any residual uncured resin; (v) drying with compressed air; and (vi) post UV-cure. After a short exposure (below 10 minutes) under the weak UV lamp, to solidify the outer surface, the parts are left under natural light for a number of hours; this slowly continues the curing process.

To fabricate the 90° OAPM, the Formlabs' Form 3 printer is used (costing approximately \$3,500). This SLA printer employs a 250 mW UV laser for illuminating (at normal incidence) and curing the resin. The Form 3 is capable of producing parts with high resolution (quoted by the manufacturer as $20\text{ }\mu\text{m} \times 20\text{ }\mu\text{m} \times 20\text{ }\mu\text{m}$ [10]) and relatively large build volume ($145\text{ mm} \times 145\text{ mm} \times 185\text{ mm}$ [10]). These attributes make it suitable for fabricating the 90° OAPM.

Formlabs's Grey resin is used, as it is suitable for applications that require a smooth surface finish and high accuracy. 3-D CAD models for the 90° OAPM are sliced to a layer-height of 20 μm using the PreForm slicer software [11]. In this case, support material is only added to non-reflecting surfaces, to avoid compromising the integrity of the mirror's curvature.

Parts produced from Form 3 require slightly different post-processing, when compared to Elegoo Mars. The finished parts are post-processed at room temperature using the following steps: (i) pre-rinse in isopropanol (IPA); (ii) mechanical support structure removal; (iii) surface cleaning using a fine brush; (iv) 15 minutes agitating in an IPA bath; (v) drying with compressed air; and (vi) post UV-cure. After a short exposure (below 10 minutes) under the weak UV lamp, to solidify the outer surface, the parts are left under natural light for a number of hours.

5.3.2 Fused Deposition Modeling

To fabricate the RAM, grid polarizer and quasi-optical housing (with an additional two-axis rotating platform), the Raise3D Pro 2 is employed (costing approximately \$4,000). This FDM 3-D printer has a maximum build volume of 280 mm $\times$ 305 mm $\times$ 300 mm [12], which makes it ideal for manufacturing large single-piece components. With its 400 μm diameter nozzle, it can reliably achieve a preferred layer height of 200 μm . 3-D CAD models for these components are sliced for printing using the IdeaMaker slicer software [13].

For the grid polarizer and quasi-optical housing, Raise 3D generic white PLA is used, while Proto-pasta conductive PLA is used for the RAM. In principle, while both the RAM and the quasi-optical housing can be co-printed (using dual extrusion), to reduce material waste, the RAM is printed separately into zig-saw bespoke tile pieces and snapped into place.

5.3.3 Metallization

The 3-D printed rectangular horn antennas and 90° OAPM are commercially coated. First, a thin electroless-plated nickel seed layer is deposited, followed by a 50 μm thick electroplated layer of copper (corresponding to 287 skin depths at the worst-case G-band frequency of 140 GHz). Finally, polishing (OAPM only) and an anti-tarnishing post-processing treatment is applied.

The grid polarizer is too thin for conventional plating. Initially, application of 842AR Super Shield™ silver conductive liquid coating (one-part durable acrylic lacquer, pigmented with an extremely conductive silver flake), having a quoted resistivity of 110 $\mu\Omega \cdot \text{cm}$, was investigated [14]. However, due to its high viscosity, the coating was uneven. An alternative could be to use the Jet Metal™ silver spray metallization process, which is reported to have a thickness-dependent volume resistivity that matches pure bulk silver of 1.59 $\mu\Omega \cdot \text{cm}$, with 2.5 to 3 μm thick layers [15]. Instead, gold leaf gilding was chosen, with bulk gold having a volume resistivity at DC of approximately 2 $\mu\Omega \cdot \text{cm}$; this material is preferred, as both silver and copper tarnish over time.

The gold leaf thickness is approximately 0.12 μm and so multiple layers are needed. The grid polarizer is pre-heated on 60°C hotplate and coated with three layers of gold leaf (corresponding to 1.7 skin depths at 140 GHz), using low-viscosity acrylic gilding glue.

5.3.4 Antenna and Mirror Weight Comparison

Mass is of great importance for some applications and so it is interesting to compare the weight of the various commercial counterparts with their 3-D printed modified replicas. To this end, all the components are weighed, and the results are given in Table 5.2. It can be clearly seen that additive manufacturing using polymer-based 3-D printing and metal plating offers a considerable weight saving; with the modified replicas being less than a third the weight of their commercial counterparts.

Table 5.2: Weight comparison for commercial counterpart reference and 3-D printed and plated modified replicas.

Component	Average Mass (g)		
	Commercial Counterpart	Modified Replica	Mass Ratio
Horn Antenna	12.5 (copper)	3.5	3.6
Parabolic Mirror	499 (aluminium)	130	3.8

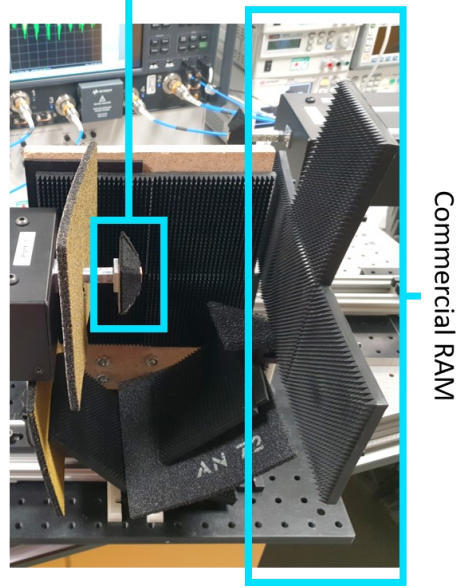
5.4 G-band Antenna and Lens Measurement

All electromagnetic measurements were undertaken within the Department of Electromagnetic and Electrochemical Technologies at the U.K.'s National Physical Laboratory, using their Keysight Technologies PNA-X N5247B vector network analyzer – with VDI WR-5.1 frequency extender heads [16] – and the 1024 pixel TeraSense TERA-1024 (sub-)THz imaging camera [17] from Imperial College London.

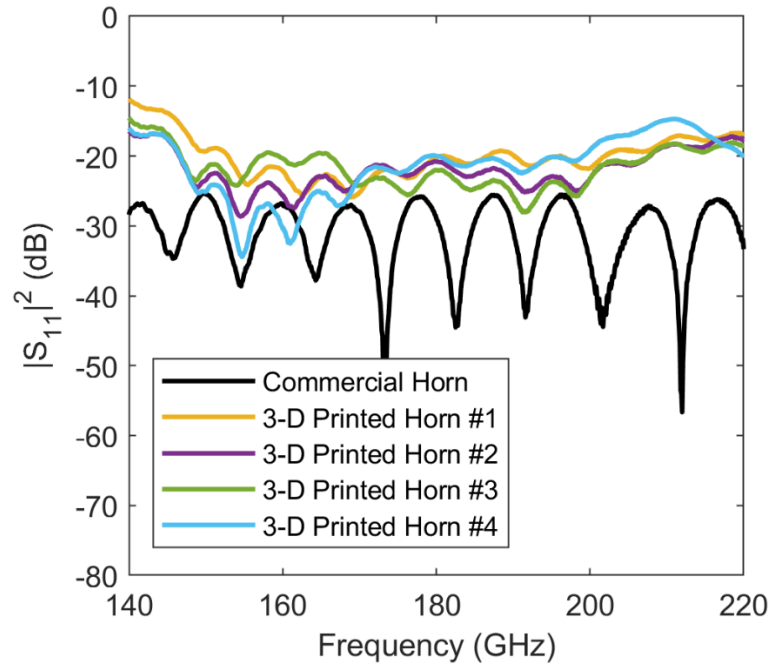
5.4.1 Horn Antenna

A semi-open anechoic environment is constructed, primarily using commercial 100 mm × 100 mm tessellating pyramidal-array RAM tiles, positioned in strategic locations (e.g., antenna boresight and above the quasi-optical breadboard). These tiles are made from solid carbon-loaded polypropylene plastic [18], manufactured by Thomas-Keating Ltd., demonstrating good measured performance from 20 to 180 GHz [19] and recommended for operation from below 50 GHz to above 1 THz [20]. In addition, a 6 mm thick flexible foam sheet broadband microwave absorber was cut to surround other strategic locations (e.g., surrounding the back of the horn antennas and covering the metal frequency extension head boxes), as shown in Figure 5.11(a). This RAM is Eccosorb® AN-72, made from multilayer carbon-loaded polyurethane foam, designed to reflect less than 17 dB of normal incident energy above 20 GHz, relative to a metal plate [21].

3-D Printed Horn Antenna
Surrounded by Commercial RAM



(a)

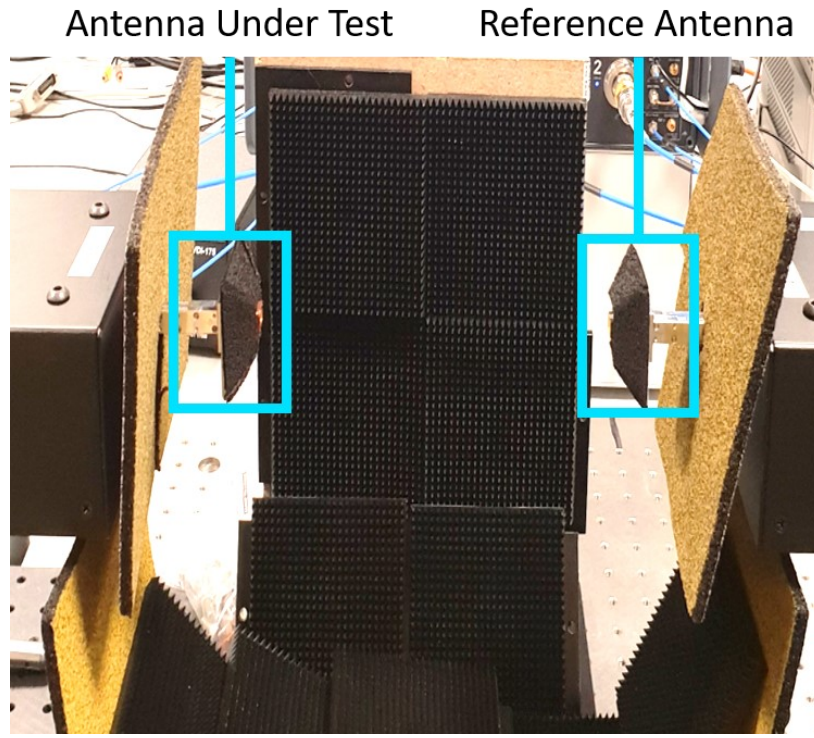


(b)

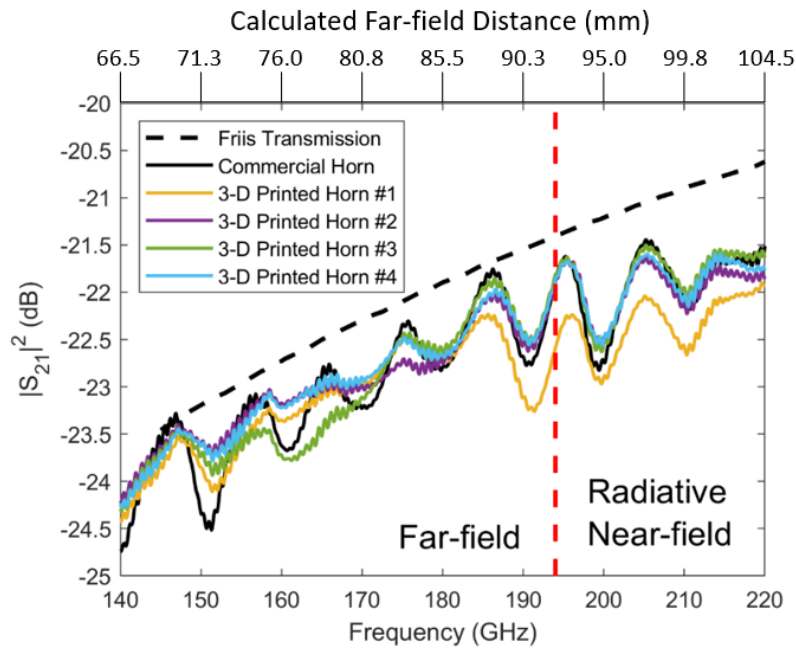
Figure 5.11: Reflection measurement setup and results for the rectangular horn antennas: (a) test setup showing the ruggedized 3-D printed modified replica antenna surrounded by commercial RAM; and (b) measured reflection coefficients for the commercial and replica antennas.

With the commercial counterpart and ruggedized 3-D printed modified replica antennas facing the RAM, their input return loss are measured, as shown in Figure 5.11(a), with the results shown in Figure 5.11(b). Across the whole of G-band, the commercial reference antenna has a return loss better than 25 dB. With the ruggedized 3-D printed antennas, the worst-case return loss is 12 dB across the whole of G-band, and approximately 20 dB from 150 to 200 GHz; demonstrating good performance with this first test scenario. The discrepancy in performance can be attributed to the lower manufacturing tolerance of the flange that has been fabricated with the low-cost MSLA 3-D printer. The low frequency Fabry-Pérot ripples (with frequency period $\Delta f \sim 20$ GHz) corresponding to the antenna flare length ($c/\Delta f \sim 14.5$ mm).

To evaluate the transmission performance, each antenna is connected in turn to Port 1, while facing a reference commercial antenna connected to Port 2; the line-of-sight aperture-to-aperture separation distance $R_s = 185$ mm. During the measurements, the facing horn-horn antenna pair is surrounded with commercial RAM, as shown in Figure 5.12(a). A commercial horn-horn antenna pair is also tested as a reference.



(a)



(b)

Figure 5.12: Transmission measurement setup and results for the rectangular horn antennas: (a) test setup showing the commercial horn antennas surrounded by commercial RAM; and (b) measured transmission coefficients for the commercial counterpart and replica antennas (1 GHz running averaged) The red dashed line shows the boundary between far field and radiative near field.

The measured insertion loss between waveguide ports is shown in Figure 5.12(b). Predictions are also included using textbook calculations for Friis transmission path loss, with quoted antenna gain values from 145 to 220 GHz [2]. It is important to note that the Friis transmission equation is only valid in the combined far-field for both antennas. As a result, the measured transmissions loss can only be compared up to 194 GHz, corresponding to the combined far-field distance of 185 mm ($92.5 \text{ mm} \times 2$). Figure 5.12(b) shows approximately 0.5 dB difference between the measured trend and predictions. Note that while the low frequency Fabry-Pérot ripples are attributed to antenna flare length, the high frequency Fabry-Pérot ripples (with frequency periods of $\Delta f \sim 1.2, 1.4, 1.6 \text{ GHz}$, $\pm 0.1 \text{ GHz}$) corresponds to the impedance mismatch reflections between both antennas. For example, between the smallest aperture size ($\Delta f \sim 1.4 \text{ GHz}$ gives $c/\Delta f \sim 214 \text{ mm}$) and largest aperture size ($\Delta f \sim 1.6 \text{ GHz}$ gives $c/\Delta f \sim 185 \text{ mm}$) for the flares of both antennas. One possible explanation for $\Delta f \sim 1.2$ could be the very weak reflections between the RAM covering the metal boxes of both frequency extension heads ($c/\Delta f \sim 250 \text{ mm}$).

Using the ‘primary gain test method’ with measured reference insertion loss $-10 \log_{10} |S_{21}|_R^2$, for the two identical commercial reference horn antennas, the reference boresight antenna gain G_R can be calculated by rearranging the Friis transmission path loss equation as [22]:

$$G_R = \left(\frac{4\pi R}{\lambda_0} \right) |S_{21}|_R \quad (5.4)$$

From (5.4), the boresight gain for the 3-D printed horn antennas under test G_T can be determined using the Friis transmission path loss equation:

$$G_T = \left(\frac{4\pi R}{\lambda_0} \right)^2 \frac{|S_{21}|_T^2}{G_R} = \left(\frac{4\pi R}{\lambda_0} \right) \frac{|S_{21}|_T^2}{|S_{21}|_R} \quad (5.5)$$

where $-10 \log_{10} |S_{21}|_T^2$ is the insertion loss with both the antenna under test and the reference antenna. The extracted boresight gains for the antennas under test are shown in Figure 5.13. The gain calculated by the manufacturer of the commercial horn antenna [2], using U.S. Naval Research Laboratory (NRL) Report 4433 (with an accuracy of ± 0.3 dB) [22], is also included for comparison. Figure 5.13 shows only approximately 0.25 dB difference between the extracted trend and the quoted gains below 194 GHz.

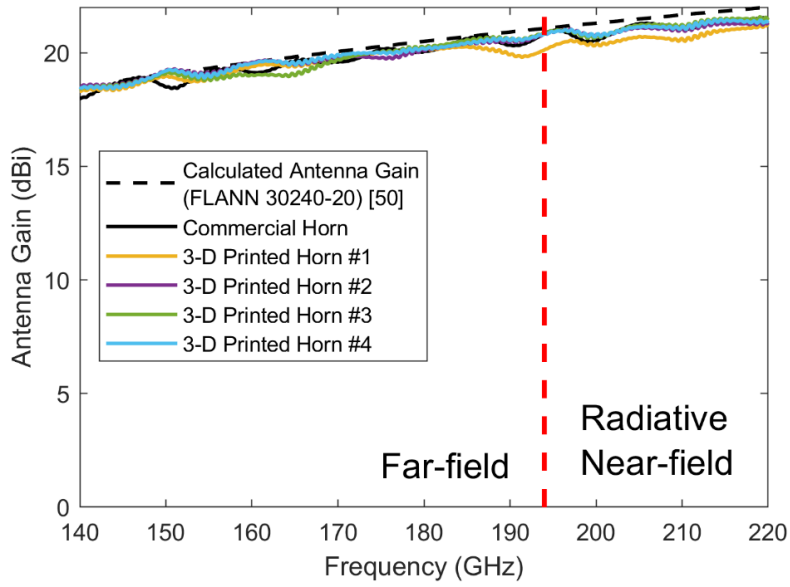


Figure 5.13: Extracted boresight antenna gains for the G-band commercial counterpart and ruggedized 3-D printed modified replica antennas. The red dashed line shows the boundary between the far-field and radiative near field.

In conclusion, all the 3-D printed modified replica horn antennas have comparable transmission performances with the commercial counterpart across the whole of G-band, again demonstrating excellent performance with this second test scenario.

To characterize the antenna beam profile within the far field, measurements are undertaken using the frequency extension heads (as a single-frequency source generator at 160 GHz) and the sub-THz imaging camera (having a peak in sensitivity at 160 GHz [17]), with its aperture located 80 cm from the antenna aperture, as shown in Figure 5.14.

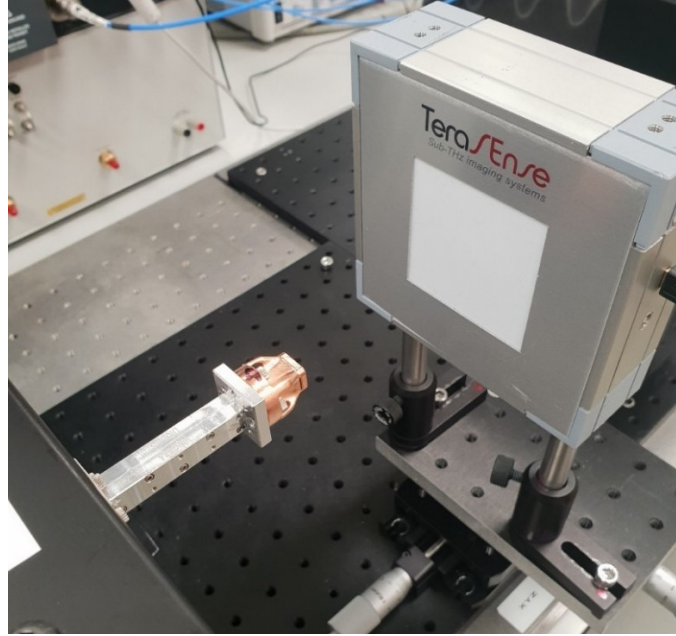


Figure 5.14: Measurement setup for spatial beam intensity profile imaging with 80 mm separation distance between antenna and camera apertures (commercial RAM is removed for this photograph).

The TeraSense camera technology can perform qualitative measurements in real-time. However, from the work on benchmarking this camera [17], it was found that the Tera-1024 TeraSense camera exhibits limited image resolution and fidelity, with significant levels of systematic spatial noise (due to faulty pixels and high pixel-to-pixel responsivity deviation). Moreover, it was also found that the TeraSense camera (with additional RAM) has an inherent limitation with its relatively high reflectance (e.g., 27% at 79 GHz) aperture [17].

With any focal plane array (FPA), image quality is dependent on both spatial resolution and pixel fidelity. The latter is a function of the number and distribution of pixels that are faulty and those with large deviations in responsivity [17]. The TeraSense camera has a specified permissible number of faulty pixels of up to 3% and the specified pixel-to-pixel responsivity deviation from the mean value is within 20% [17].

Nevertheless, at 76 and 92 GHz, it has been shown that the spatial beam profile for a W-band rectangular horn antenna can be displayed by averaging-out

pixel-related issues, using post image processing with the MATLAB ‘disk smooth’ filter [17]. For this reason, the same process is applied here.

The measured and post image processed spatial beam intensity 2-D profiles (normalized to their spatial peak intensity) for the commercial counterpart and ruggedized 3-D printed modified replica horn antennas are shown in Figure 5.15(b) to (f). EM simulated results are also shown in Figure 5.15(a), as a reference.

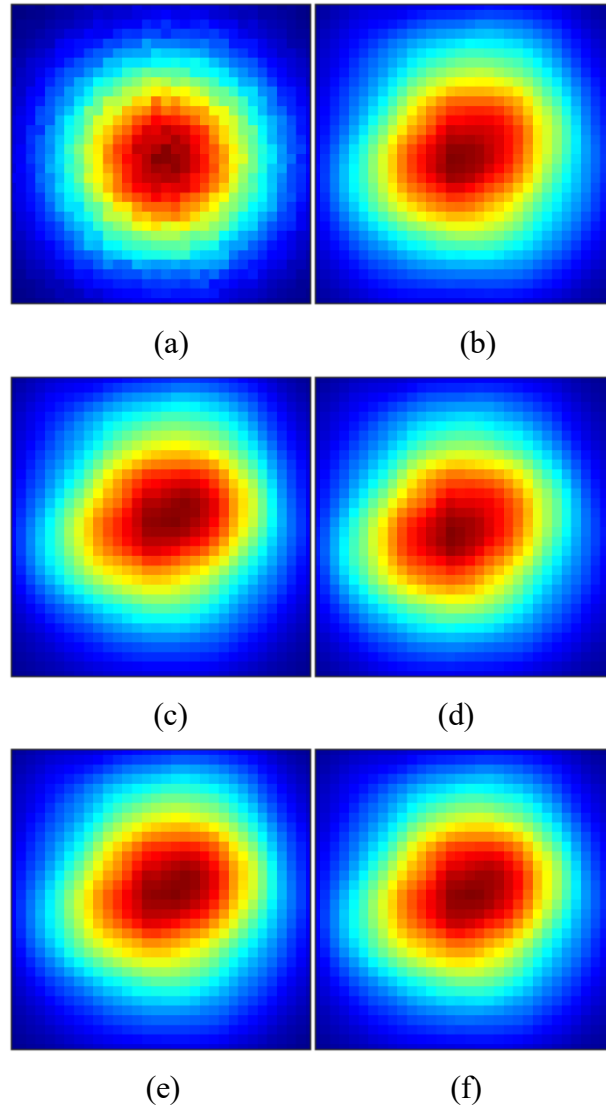
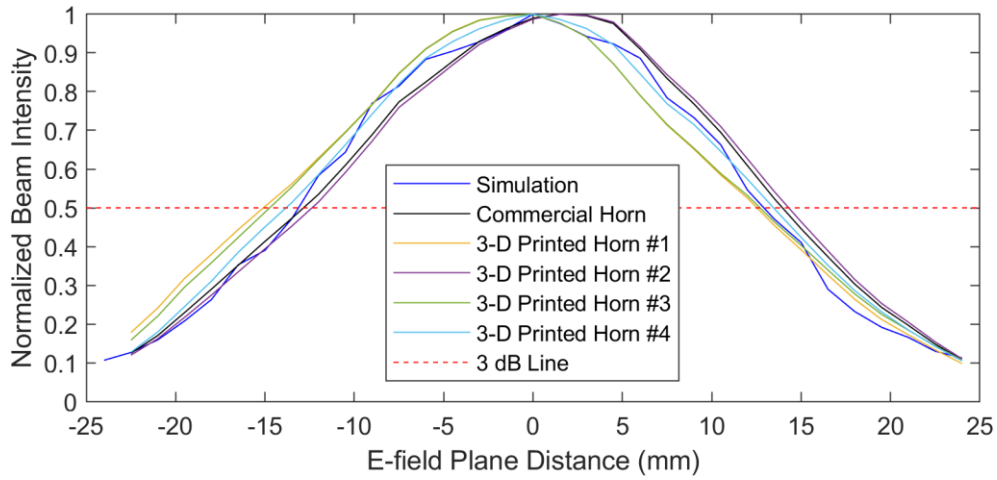
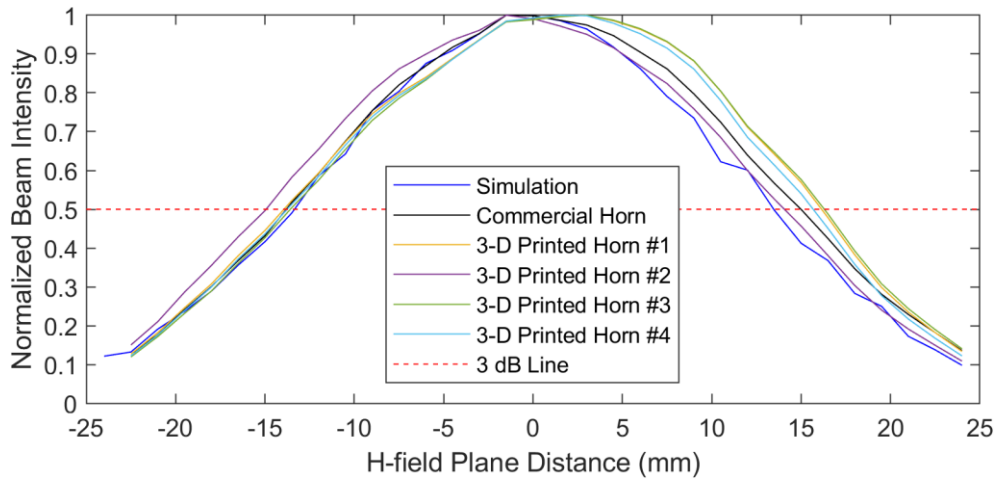


Figure 5.15: Measured normalized spatial beam intensity 2-D profiles for the G-band horn antennas with 80 mm separation distance between antenna and camera apertures: (a) HFSS simulation; (b) commercial horn; (c) Horn #1; (d) Horn #2; (e) Horn #3; and (f) Horn #4

Figure 5.15 clearly shows that the beam profiles for the commercial counterpart and 3-D printed replica horn antennas have almost identical shapes and sizes. When compared to the circular EM simulated profile, all the measured profiles are slightly skewed. Possible reasons for this profile distortion include a combination of angular misalignment, pixel-related issues and/or standing waves [17].



(a)



(b)

Figure 5.16: Cross-sectional beam intensity profiles for the G-band horn antennas extracted from Figure 5.15: (a) E-field plane; and (b) H-field plane.

Figure 5.16 shows the measured cross-sectional E- and H-field plane beam intensity profiles extracted from Figure 5.15. It can be seen that the antenna-camera spatial alignment has shifted between each antenna measurement. It is believed that this may be a combination of small variations in torque applied to the screws holding the flanges together and slight movement in the frequency extension head. Nevertheless, it can be seen that all the E- and H-field plane beam intensity profiles for the 3-D printed antennas match both the commercial counterpart and the EM simulations very well, again demonstrating excellent performance with this third test scenario.

Table 5.3: Extracted 3 dB beamwidths for the G-band horn antennas.

Horn Antennas	3 dB Beamwidth (°)	
	E-field Plane	H-field Plane
Calculated [2]	19.1	18.8
HFSS Simulation	18.5	19.1
Commercial Horn	19.0	20.4
3-D Printed Horn #1	19.6	21.3
3-D Printed Horn #2	19.4	20.5
3-D Printed Horn #3	19.5	21.2
3-D Printed Horn #4	19.3	20.8

Table 5.3 gives the 3 dB beamwidths calculated and quoted by the manufacturer of the commercial horn antenna [2] and also the measured values extracted from Figure 5.16. The difference between quoted and measured 3 dB beamwidths for E- and H-field planes are 0.1° and 1.6° , respectively, for the commercial horn antenna; the latter being relatively large, due to the skewed 2-D beam profiles. The worst-case differences in measured 3 dB beamwidths, between the modified replica and commercial counterpart antennas, are 0.6° and 0.9° for the E- and H- planes, respectively. Note that the small discrepancies in repeatability are due in part to the many limitations of the TeraSense camera (e.g., spatial resolution and pixel sensitivity); being difficult to accurately quantify.

In summary, from the measured performances of the ruggedized 3-D printed horn antennas, it can be concluded that they have excellent mm-wave performance for replicating their commercial counterpart.

5.4.2 Dielectric Lenses

The maximum output power from the VDI WR-5.1 frequency extension head is only 6 dBm [23]. With the 90° OAPM, the two-sigma (i.e., $1/e^2$ normalized power) boundary for the collimated Gaussian beam (containing 95.45% of the beam power) projecting onto the TeraSense camera aperture cannot be determined, due to the poor signal-to-noise ratio at the FPA. Nevertheless, from simulation, the $1/e^2$ beamwidth is less than 50 mm and, as a result, the 60 mm diameter lenses are sufficiently large for this measurement.

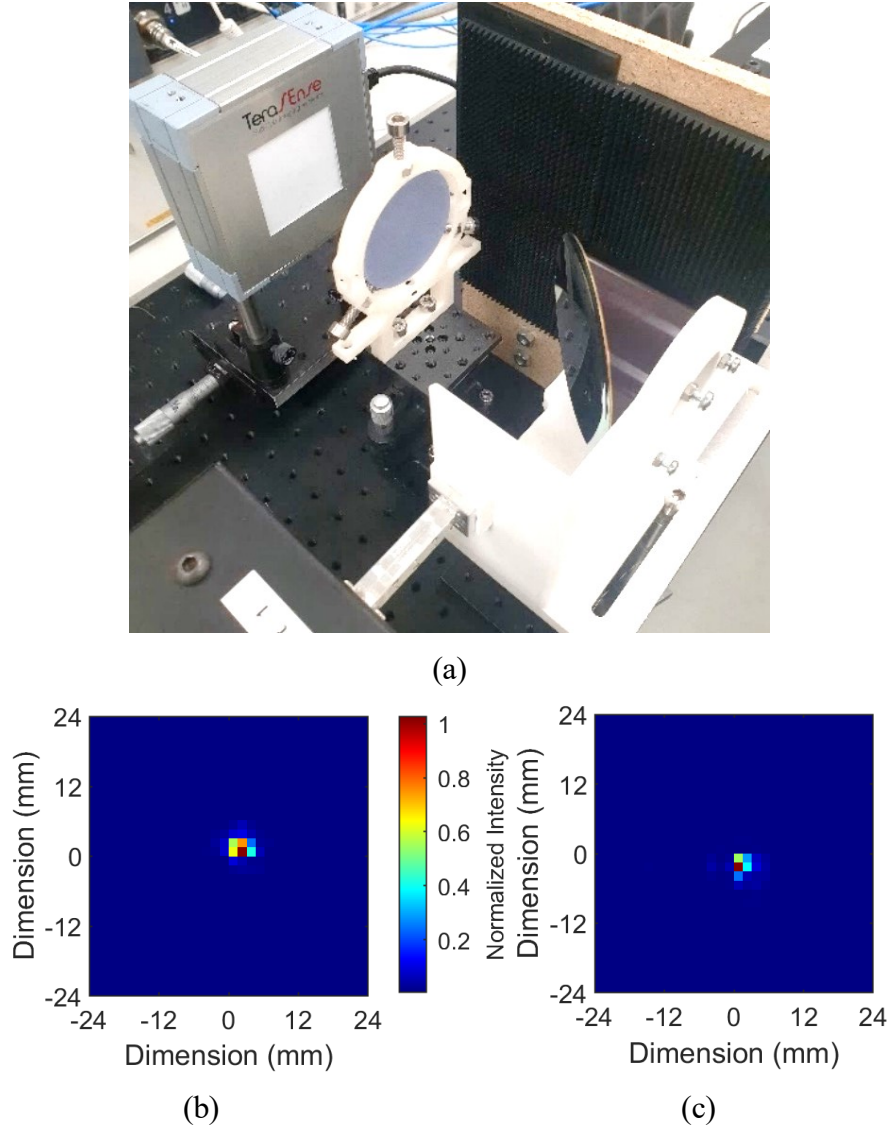


Figure 5.17: Measured spatial beam intensity 2-D profile images for the hyperbolic aspherical plano-convex lenses: (a) measurement setup; (b) conventional lens focal point image; and (c) Fresnel lens focal point image.

To test the 3-D printed lenses, a lens holder and horn-mirror alignment frame are FDM 3-D printed, as shown in Figure 5.17(a). Here commercial horn antenna and OAPM are employed. The collimated beam is focused by the lens and its focal point is found by moving the camera along the propagation z -axis, as shown in Figure 5.17. It is important to note that this method is only qualitative, due to the dominant standing-wave between the TeraSense camera aperture and the lens [17].

With the conventional hyperbolic aspherical plano-convex lens, the approximate focal length is 73 mm, which is very close to the designed value of 76.2 mm. The associated focused beam spot image is shown in Figure 5.17(b). Here, only one pixel displays the unity value of normalized intensity (shown as a red pixel). The pitch of the pixels in the FPA is 1.5 mm, while the size of each individual detector within each pixel area is unknown. For this reason, it is not possible to determine the exact spot size.

With the Fresnel lens, the approximate focal length is 61 mm and the corresponding focused beam spot image is shown in Figure 5.17(c). Again, only one pixel displays the unity value of normalized intensity (shown as a red pixel). The deviation in focal length, from its designed value of 76.2 mm, is much bigger with the Fresnel lens. Possible reasons for this are the non-optimized design and significant warpage. Nevertheless, both ultra-low cost lenses demonstrate their capability for focusing a collimating quasi-optical beam.

5.5 G-Band Mirror, Polarizer and RAM Measurement

The 3-D printed quasi-optical housing is utilized to evaluate the performances of the 90° OAPM, grid polarizer and RAM. The ‘plug and play’ housing can accommodate a mix of commercial and 3-D printed components and their configurations, as a quasi-optical subsystem assembly. Figure 5.18fig (a) illustrates a typical configuration for the quasi-optical housing. Custom-built 3-D printed mechanical interface adapters are used between the horn antennas and housing. Photographs for the complete integrated subsystem assembly are shown in: (i) Figure 5.6, using commercial antennas and mirrors; (ii) Figure 5.18(b), using all 3-D printed quasi-optical components (antennas, mirrors, RAM and polarizer).

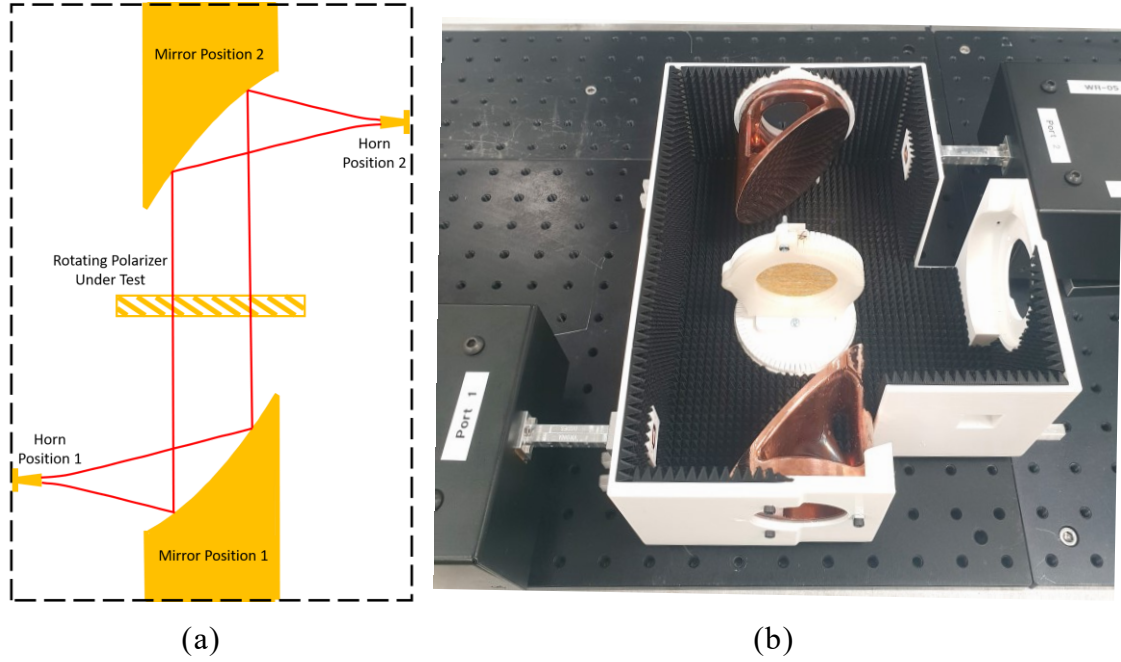


Figure 5.18: Typical configuration for the 3-D printed quasi-optical housing with HMMH arrangement having the two OAPMs rotated 180° about their common optical axis and the central two-axis rotating platform containing the grid polarizer under test: (a) illustration; (b) corresponding photograph of the complete integrated subsystem assembly realized using all 3-D printed quasi-optical components (antennas, mirrors, RAM and polarizer).

5.5.1 Parabolic Mirrors

As a reference, to evaluate the performance of the 3-D printed 90° OAPMs, the housing is first assembled using all commercial antennas and mirrors (without the rotating platform). The commercial mirror at Mirror Position 1 is replaced with its mass-reduced 3-D printed modified replica. The measured insertion loss for the commercial counterpart and its replica mirrors are shown in Figure 5.19. It can be seen that all the 3-D printed mirrors have excellent performances, with measured port-to-port insertion loss between 2.0 to 3.5 dB across G-band. By comparison, the National Metrology Institute of Germany, Physikalisch-Technische Bundesanstalt (PTB), reported an insertion loss better than 4 dB, between 50 GHz and 325 GHz, using identical commercial components and their custom-machined mechanical alignment rig (90° rotated mirror mounting) [24], [25]. This clearly shows that the 3-D printed quasi-optical housing offers very

accurate spatial alignment over the complete subsystem assembly, having a total port-to-port path length of 356 mm (corresponding to $166 \lambda_0$ (140 GHz)). Moreover, there is no noticeable difference in measurements between commercial counterpart and mass-reduced 3-D printed modified replica mirrors, which shows the choice of 3-D printing technologies (FDM for the housing and SLA for the mirrors) provides excellent dimensional accuracy for their intended purposes.

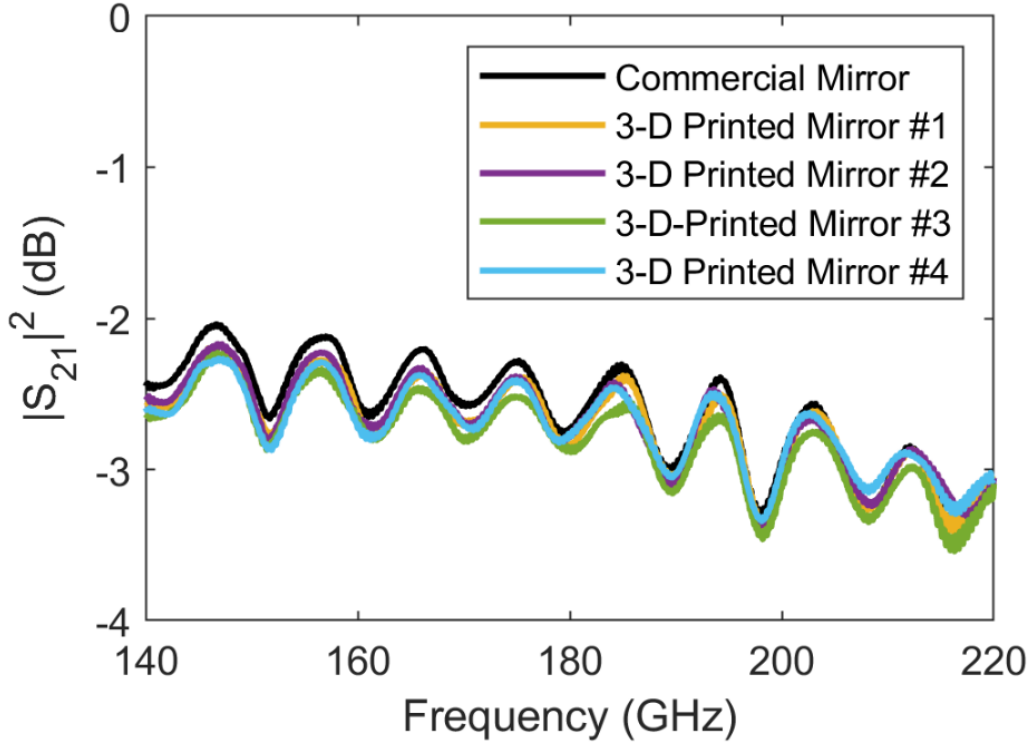


Figure 5.19: Measured port-to-port transmission coefficients for the commercial counterpart and mass-reduced 3-D printed modified replica 90° OAPMs (2 GHz running averaged with 10 data points each side).

In addition, to evaluate the quasi-optical subsystem assembly employing either all commercial counterpart antennas and mirrors (Figure 5.6, without the rotating platform) or all 3-D printed modified replicas (Figure 5.18(b), without the rotating platform), the S-parameter measurements for both scenarios are shown in Figure 5.20. It can be seen that all 3-D printed subsystem shows excellent performance. This clearly demonstrates that a complete quasi-optical subsystem can be entirely manufactured out of 3-D printed components, without significantly sacrificing performance, at G-band.

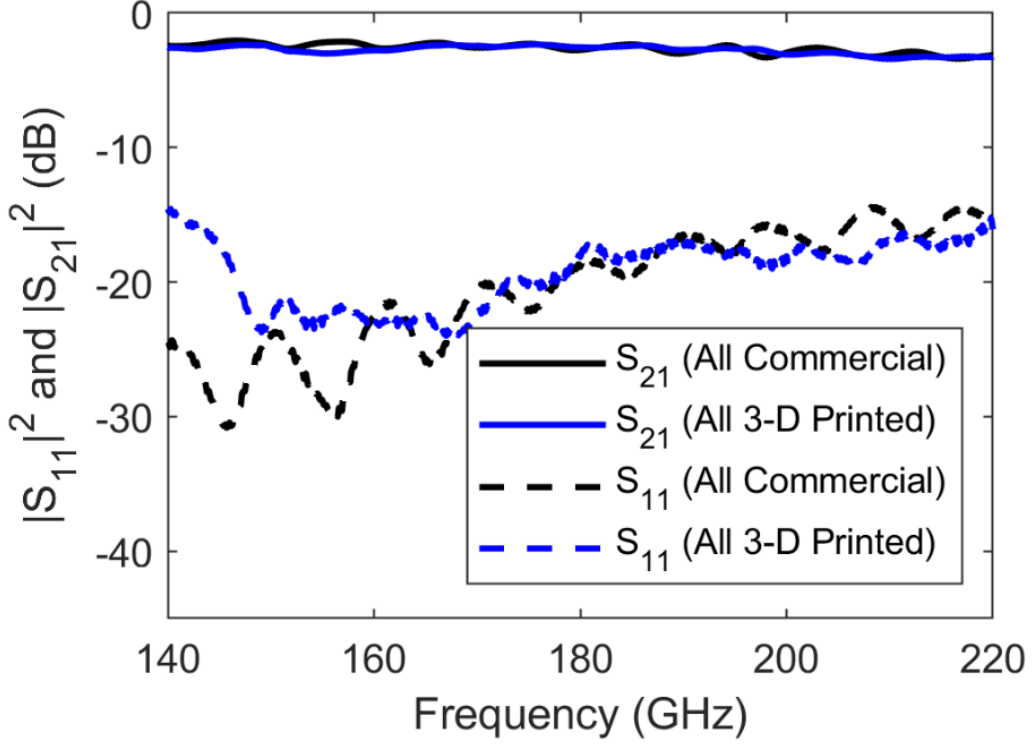


Figure 5.20: Measured S-parameters for the quasi-optical subsystem assembly employing either all commercial counterpart antennas and mirrors or all 3-D printed modified replicas, over a path length of 356 mm (2 GHz running averaged)

5.5.2 Grid Polarizer

To evaluate the 3-D printed grid polarizer, the housing is fitted with the rotating platform (having 3-D printed alignment markers at every 5° intervals of rotation) and only commercial antennas and mirrors are employed. Figure 5.6 shows the in-situ measurement setup of the linear polarizer.

The embedded polarizer subsystem (denoted by subscript $p-s$) S-parameter measurements S_{11p-s} and S_{21p-s} are recorded for every 5° of polarizer rotation angle θ_p , ranging from 0° (horizontal grid lines) to 90° (vertical grid lines). With the vertically polarized rectangular horn antennas, at 160 GHz, the insertion loss is at a minimum with $\theta_p = 0$ and maximum with $\theta_p = 70^\circ$ (and not at the expected 90° , possibly due to stress-induced mechanical distortion of the grid), as shown in Figure 5.21(a). At 160 GHz, Figure 5.21(b) shows the measured rotation angle dependency for the embedded polarizer power reflectance $\mathcal{R}_{p-s}(\theta_p) =$

$|S_{11p-s}(\theta_p)|^2 \xrightarrow{\text{lossless}} \sin^2 \theta_p$, transmittance $\mathcal{T}_{p-s}(\theta_p) = |S_{21p-s}(\theta_p)|^2 \xrightarrow{\text{lossless}} \cos^2 \theta_p$ and absorptance [26] (also known as power loss factor):

$$\mathcal{A}_{p-s}(\theta_p) = 1 - |S_{11p-s}(\theta_p)|^2 - |S_{21p-s}(\theta_p)|^2 \quad (5.6)$$

Using the principle of conservation of energy, given in (5.6), the solid lines in Figure 5.21(b) show the semi-analytical modeling for power reflectance and transmittance, based on the extracted absorptance.

The respective polarizer angle-frequency surface plots for power reflectance, transmittance and absorptance are shown in Figure 5.21(c), (d) and (e).

The embedded absorptance (for the complete subsystem assembly) is between 57% and 90%, from 140 to 187 GHz, indicating poor power efficiency due to ohmic and radiative (diffraction) losses. In practice, the de-embedded absorptance will be better. Since $|S_{21p-s}|^2$ includes the insertion loss of the quasi-optical housing (without the polarizer), the de-embedded polarizer power transmittance $\mathcal{T}_p$ can be approximated using power propagation modeling [26] from:

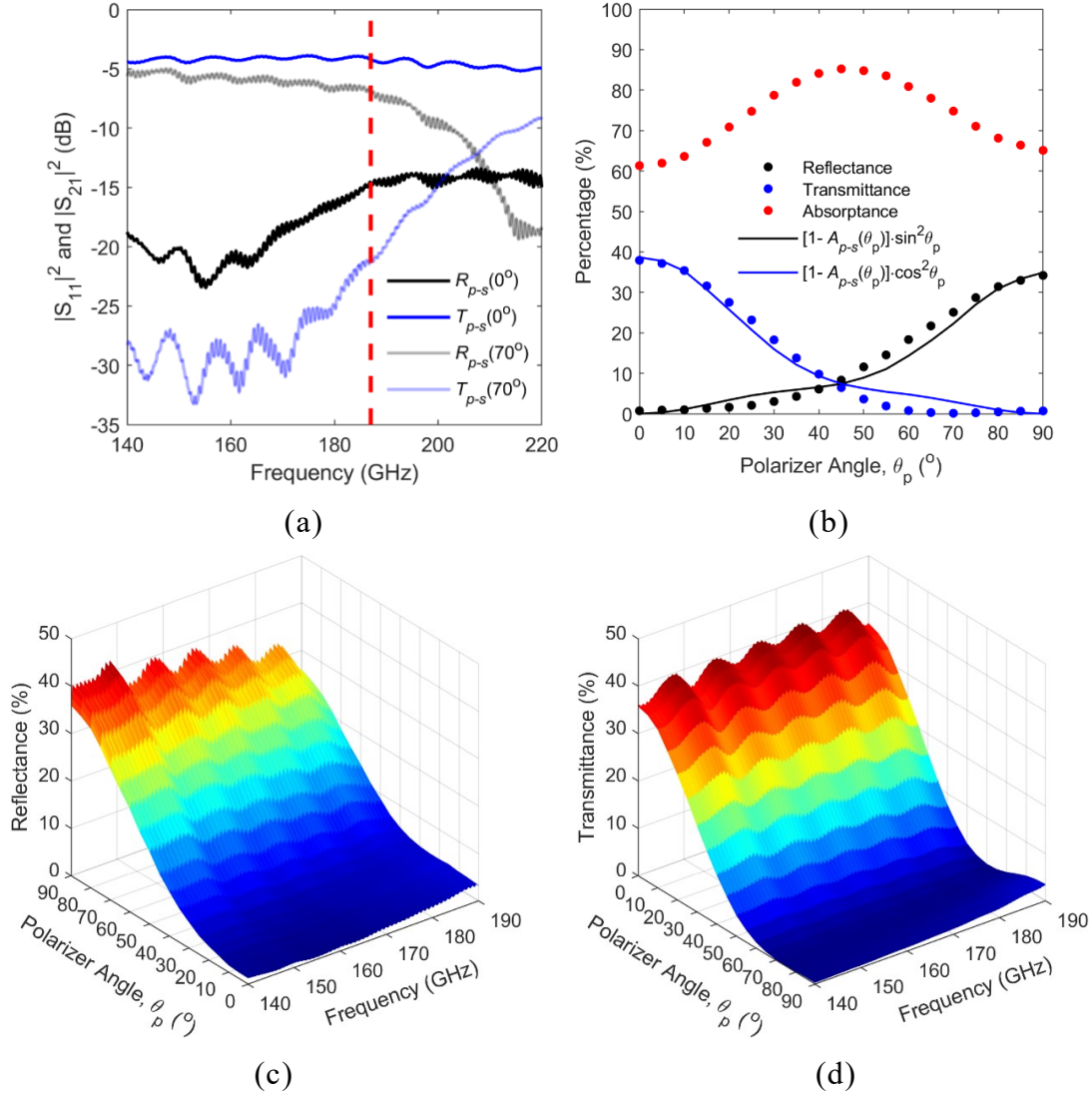
$$\mathcal{T}_p(\theta_p) \simeq \frac{|S_{21p-s}(\theta_p)|^2}{|S_{21}|^2} \quad (5.7)$$

where $|S_{21}|^2$ is the measured insertion loss of the quasi-optical housing without the polarizer, given by the solid black line in Figure 5.20. The degree of polarization (or polarization efficiency) P and extinction ratio E_p are calculated from:

$$P = \frac{E_p - 1}{E_p + 1} \quad \text{and} \quad E_p = \frac{\mathcal{T}_p(0^\circ)}{\mathcal{T}_p(70^\circ)} \approx \frac{|S_{21p-s}(0^\circ)|^2}{|S_{21p-s}(70^\circ)|^2} \quad (5.8)$$

The polarization efficiency and extinction ratio across G-band are shown in Figure 5.21(f). With the 3-D printed design, the theoretical TE upper cut-off frequency is 187 GHz. At this frequency, the polarization efficiency and extinction ratios are 96% and 48%, respectively. From 140 to 180 GHz, these respective values are better than 98% and 118. This compares favorably with commercial free-standing wire grid polarizers, used for ultra-broadband (sub-)millimeter wave applications, having low to medium values of extinction ratio

when to compared to other polarizer technologies. For example, the PW010-040-050 (from PureWavePolarizers Ltd., U.K.) is a 50 mm diameter (34 mm aperture) grid polarizer that uses 10 μm diameter tungsten wires, spaced 40 μm apart and, at 150 GHz, has normalized transmission values of $\bar{T}_p(0^\circ) \simeq 99.9\%$ and $\bar{T}_p(90^\circ) \simeq 0.3\%$, respectively [27]; giving polarization efficiency and extinction ratio values of 99.4% and $E_p \approx 333$ at 150 GHz.



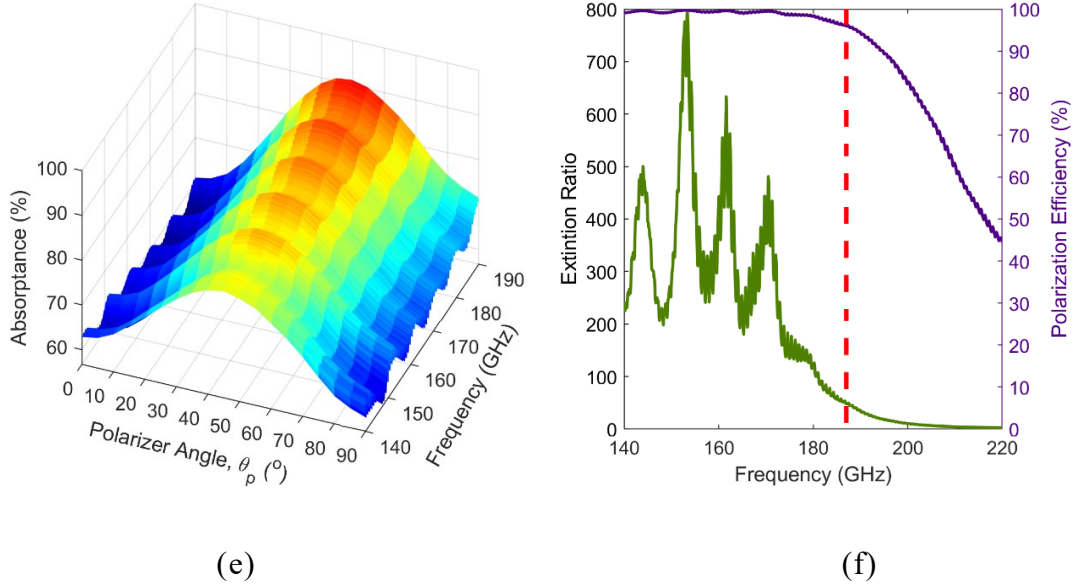


Figure 5.21: Measurement results for the grid polarizer: (a) embedded S-parameters, with transmittance (dB) in blue and reflectance (dB) in black – darker and lighter colours are for 0° and 70° polarizer angles, respectively; (b) embedded polarizer power reflectance, transmittance and absorptance against polarizer angle at 160 GHz; embedded polarizer angle-frequency surface plot for (c) reflectance, (d) transmittance and (e) absorptance; and (f) polarization efficiency and extinction ratio performance. In (a) to (f), there is a 2 GHz running averaged with 10 data points each side and the red dashed line shows the TE upper cut-off frequency.

Note that the in-band high frequency Fabry-Pérot ripples, seen in Figure 5.21(a) and (f), (with frequency period $\Delta f \sim 1.7$ GHz) corresponds to reflections between the grid polarizer and both antenna apertures ($c/\Delta f \sim 178$ mm).

The results show that with a low-cost FDM 3-D printer, even with its poor resolution, a usable grid polarizer can be fabricated. Moreover, while this first non-optimized proof-of-principle demonstrator exhibits poor power efficiency, it performs its primary function, with sufficient extinction ratio below its 187 GHz TE upper cut-off frequency.

5.5.3 Radiation Absorbent Material

To evaluate the performance of the 3-D printed RAM, a tile was placed against a flat copper back plate and illuminated at normal incidence with a collimated beam. The RAM tile and back plate are mounted on the rotational platform within

the 3-D printed quasi-optical housing, as shown in Figure 5.22(a). It is important to note that the commercial horn antenna used is not ideal, having a finite return loss. As a result, the dynamic range for this measurement scenario is limited by the performance of the commercial horn antenna.

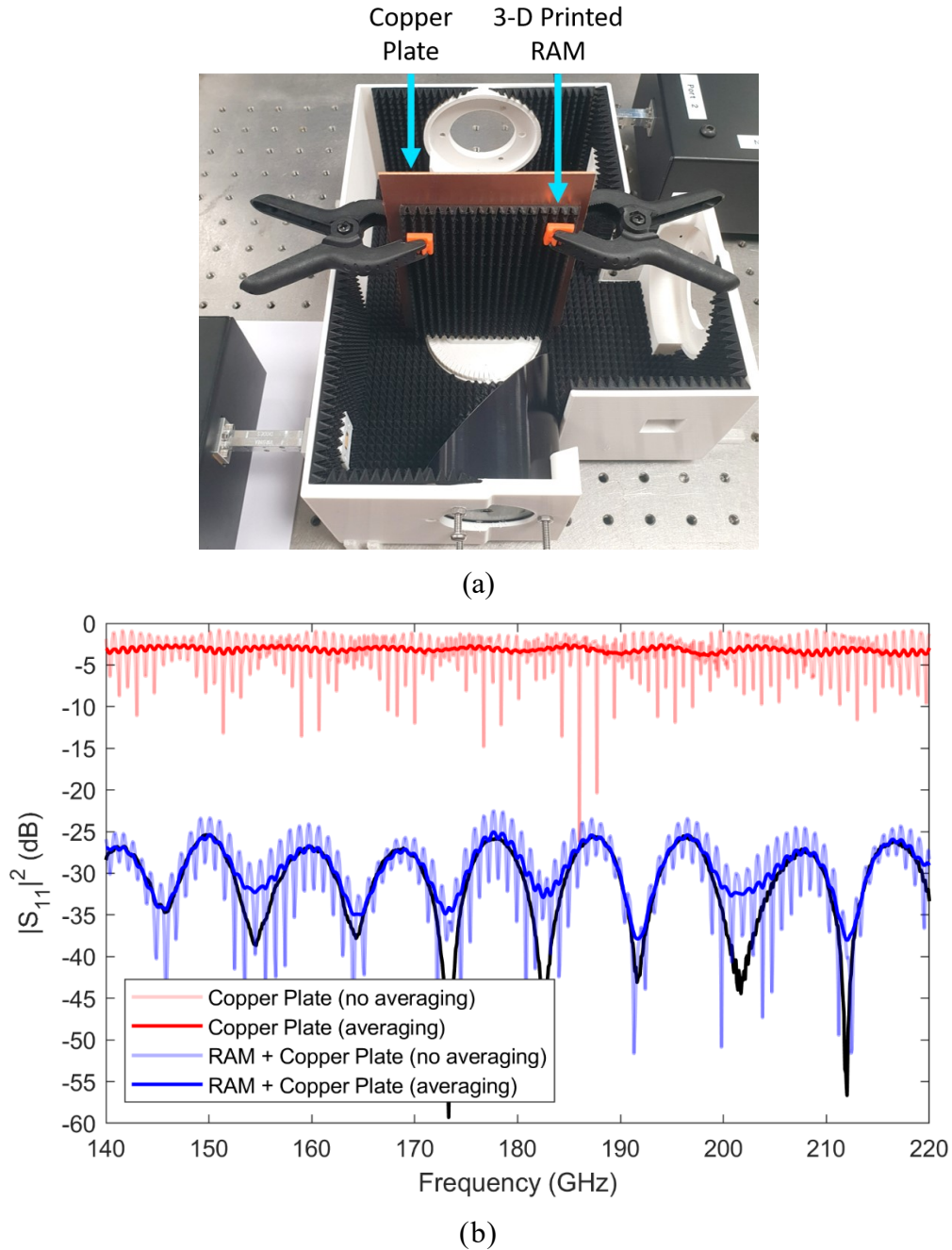


Figure 5.22: Measurement of the 3-D printed RAM (with copper back plate) exposed to a collimated beam at normal incidence: (a) test setup with commercial horn and mirror; (b) reflection coefficients (commercial horn measurement from Figure 5.11 is shown by the black curve). With the darker lines there is a 1 GHz running averaging

The measured return loss for the copper back plate and 3-D printed RAM (against the copper back plate) is shown in Figure 5.22(b). With the former, the copper plate reflects most of the power, and the resulting G-band average return loss is 2.98 dB. With a perfect lossless system, the return loss from the copper plate should be 0 dB. However, the band-average return loss is consistent (within 0.37 dB) with the previous port-to-port measurement for the horn-mirror-mirror-horn insertion loss of 2.61 dB. When RAM is added, the return loss decreases below 25 dB, which is near the lower end of the dynamic range limitation for the scenario. This clearly shows that the 3-D printed RAM is capable of suppressing normal incidence reflections (i.e., two-way transmission through the RAM) by at least 21 dB. While absorption is the primary mechanism for this high value of suppression, scattering will also contribute to some degree. Note that the high frequency Fabry-Pérot ripples correspond to the weak reflections between the RAM-covered copper back plate and the antenna; the amplitude of these ripples will reduce with an increase in the base thickness of the RAM.

To observe the performance of the RAM for a diverging incident beam, it is placed mid-way between commercial horn antennas, facing one another, shown in Figure 5.23(a). The antenna at Port 2 is facing the front pyramidal side of the RAM, while the antenna at Port 1 is facing the back flat side of the RAM, with measured return loss for both ports shown in Figure 5.23(b).

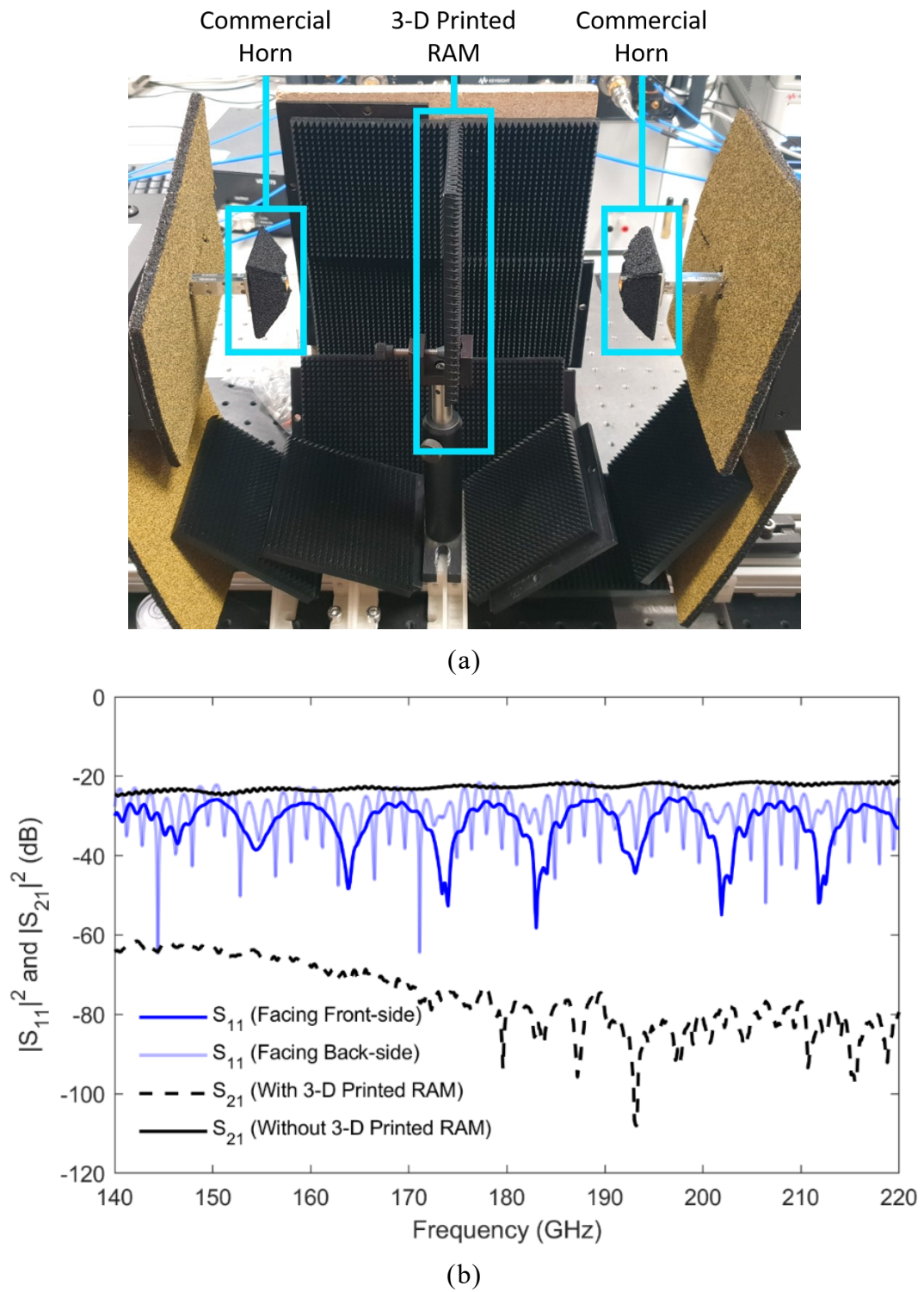


Figure 5.23: Horn-RAM-horn measurements (no averaging) with 3-D printed RAM (without copper back plate) inserted mid-way between antennas: (a) test setup with commercial horn antennas; (b) S-parameters.

Figure 5.23(b) shows that without the RAM, the worst-case insertion loss is approximately 23 dB. When the RAM is added, the worst-case insertion loss is 60 dB. This clearly shows that adding the 3-D printed RAM provides better than 37 dB of isolation. In addition, as expected, the pyramidal front side of the RAM reduces reflections more than the flat backside. With the former, the dominant low frequency Fabry-Pérot ripples are attributed to the illuminating antenna itself. With the latter, the high frequency Fabry-Pérot ripples (with frequency period $\Delta f \sim 3.2$ GHz) corresponds to the distance between the backside of the RAM and the antenna ($c/\Delta f \sim 88.5$ mm). This clearly indicates that the wave impedance transforming gradient, with the tessellating pyramidal array on the front side, allows the FDM 3-D printed RAM to be highly effective across the whole of G-band.

5.6 Conclusion

In this work five different quasi-optical component types are investigated; antennas, mirrors, RAM, polarizers and lenses. As an alternative to conventional electroplating, gold-leaf gilding is used for the polarizer. A detailed investigation is undertaken to compare the performance of the 3-D printed horn antenna, parabolic mirror and RAM with their commercial equivalents; demonstrating comparable performances. Although the FDM 3-D printed grid polarizer is limited by the minimum strip width, separation distance and metallization thickness, it shows potential for further development using a smaller nozzle or different 3-D printer technology (e.g., MSLA, SLA, RECILS). The hyperbolic aspherical plano-convex lenses both demonstrated their focusing ability. All the antennas and lenses are fabricated using one of the lowest cost 3-D printing technologies (MSLA) available on the open market.

In addition, a fully 3-D printed, RAM-lined housing with a central two-axis rotational platform is constructed. This ultra-low cost, ‘plug and play’ housing is designed to give a fast measurement setup, while minimizing misaligning losses. Its RAM lining is designed to suppress reflections due to diffraction from components under test that may cause adverse multi-path interference. Using this housing accurate two-port quasi-optical measurements has been demonstrated.

5.7 Acknowledgements

Work presented in this chapter was published in the paper: “Polymer-based 3-D printed 140-220 GHz low-cost quasi-optical components and integrated subsystem assembly” by **S.-H. Shin**, X. Shang, N. M. Ridler, and S. Lucyszyn, (*IEEE Access*, vol. 9, pp. 28020-28038, Feb. 2021) [1]. The research on this work has been performed under the supervision of Prof. S. Lucyszyn and he went over every word of the original article. Measurements at NPL were performed under the supervision and help of X. Shang and N. M. Ridler.

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6 3-D Printed Plug and Play Subsystem

In this chapter, using various 3-D printing techniques, both waveguide and quasi-optical components, are integrated to create a compact multi-channel subsystem proof-of-concept demonstrator. This work is based on published article by Shin *et al.* [1]. Here, the proposed front-end sub-THz subsystem integrates 30 individual components: 1× OAPM, 1× plano-concave lens, 1× reconfigurable flat mirror, 5× flat mirrors, 4× beamsplitters, 6× plano-convex lenses, 6× horn antennas and 6× BPFs. Identical OAPM and horn antennas were previously developed (Chapter 5 [2]), while the plano-convex lens has been redesigned for this new bespoke application. As a result, details of their design, fabrication and characterization can be found in Chapter 5. This chapter introduces concave lenses and flat mirrors, while concentrating on a novel beamsplitter, requiring custom 3-D printed test platforms. More importantly, this work demonstrates the proof-of-concept for a fully 3-D printed integrated plug and play prototype architecture.

6.1 Design

The design goal is to implement a compact front-end subsystem architecture (with details given in Subsection 6.1.4). In this section, the design of the bespoke prototype dielectric lenses, holey beamsplitter and also the test platforms and subsystem housings will be discussed. From the outset, it was decided that the complete single-piece subsystem housing would fit within the maximum build volume (280 mm × 350 mm × 300 mm) of the FDM 3-D printer (Raise Pro 2). Multiple housing parts would compromise alignment accuracy and structural integrity.

6.1.1 Dielectric Lenses

The compact nature of the architecture dictates a number of design implantation choices. For example, lenses are chosen over mirrors to focus and collimate beams, minimizing signal path complexity; this also allows a significant reduction in both volume and mass, but increases transmission losses (discussed

in subsection 6.3). Two different types of hyperbolic aspherical lenses are designed for the subsystem: a plano-convex lens and a plano-concave lens.

The main input test beam is externally collimated by a commercial antenna-OAPM pair [2]. With the OAPM, the two-sigma (i.e., $1/e^2$ normalized power) boundary for the collimated Gaussian beam (containing approximately 95% of the beam power) projected into the subsystem's main input window has an estimated $1/e^2$ beamwidth that is less than 50 mm (i.e., below the 76.2 mm diameter of the OAPMs, to minimize diffraction losses with both the external commercial OAPM and its internal 3-D printed OAPM replica [2]).

The internal beamwidth is significantly reduced to approximately 20 mm, to reduce the width of the lenses (helping to make the subsystem more compact) and, therefore, also reduces their thickness (lowering transmission losses with plano-convex lenses). At the output of the internal OAPM, a convex lens could have been chosen. However, in order to significantly reduce the transmission path length (to achieve a compact subsystem), a hyperbolic aspherical plano-concave lens is chosen. In addition, six hyperbolic aspherical plano-convex lenses are employed for the 6-channel architecture, each focusing the internal collimated beam into its corresponding channel output horn antenna.

The building material for the lenses is chosen to be the Elegoo Water Washable Rapid Resin (Ceramic Grey), as its dielectric properties are already characterized from the previous chapter: dielectric constant $\epsilon_r \approx 2.89$ and loss tangent $\tan\delta \approx 0.032$ across G-band. The hyperbolic aspherical curvatures for both plano-concave and plano-convex lenses, use the following design equation [2]:

$$z^2(r)(\epsilon_r - 1) + 2f_{pl}z(r)(\sqrt{\epsilon_r} - 1) - r^2 = 0 \quad (6.1)$$

where f_{pl} is the target focal path length, $z(r)$ is the curvature profile and r is the corresponding radial distance. Also, to verify the design, a custom MATLAB code is written for optical ray-tracing, with results shown in Figure 6.1(a). It can be seen that the plano-convex lens design can provide a perfect optical focus with the target focal path length of $f_{pl} = 50$ mm. Figure 6.1(b) shows that the plano-

concave lens is able to collimate a converging beam. In both cases, spherical aberration is not theoretically present. In practice, however, because the input source cannot be perfectly collimated, the lenses will not be able to focus/collimate perfectly. 3-D printed lenses and the key properties of the lenses are shown on Figure 6.1 and Table 6.1: Design properties for the dielectric lenses, respectively.

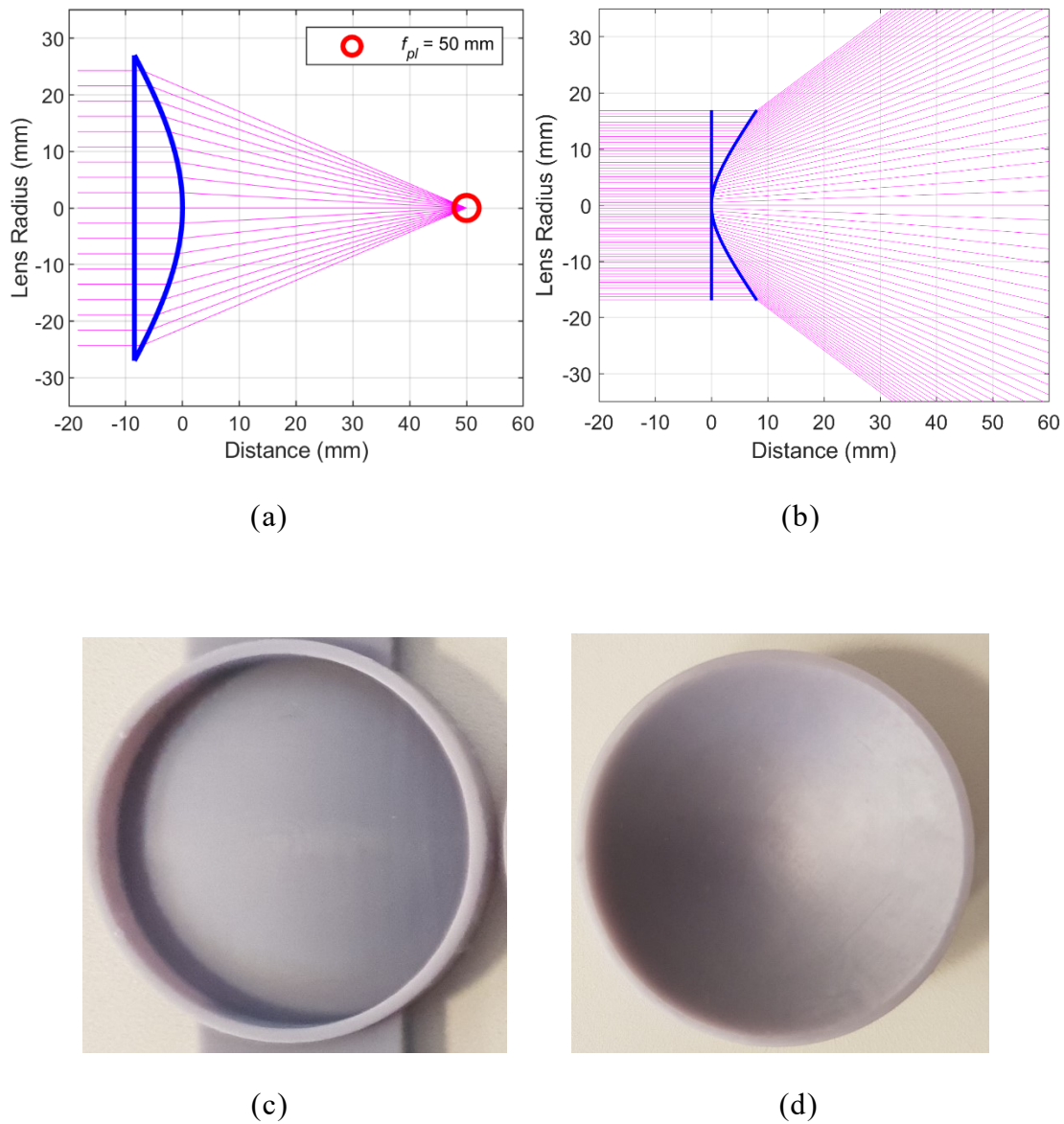


Figure 6.1: 3-D printed lenses: (a) simple ray tracing for the plano-convex lens; (b) simple ray tracing for the plano-concave lens; (c) photograph of the plano-convex lens; and (d) photograph of the plano-concave lens.

Table 6.1: Design properties for the dielectric lenses

Property	Plano-convex	Plano-concave
Center Thickness (mm)	4.589	1.000
Edge Thickness (mm)	10.00	9.070
Diameter (mm)	35.00	33.91
Rim Width (mm)	1.250	1.000
Material	Elegoo Water Washable Gray	
Refractive Index [2]	1.7 (G-band)	

6.1.2 Holey Beamsplitter

To split an incident collimated beam into two separate beams of equal power, a holey beamsplitter is developed for this application. Conventional wideband free-space beam splitting is commonly achieved using a ‘polka dot beamsplitter’ at optical frequencies (infrared and visible) [3], [4]. The polka dot beamsplitter consists of a 2-D periodic array of metalized patches (typically circular or square) that masks 50% of its glass substrate. Because the size of each patch is sufficiently larger than the maximum in-band optical wavelength, it has an excellent performance over a wide bandwidth [5].

For (sub-)THz frequencies, sub-wavelength metasurface beamsplitters have been reported, based on either a 2-D or 3-D periodic array [6]-[10]. However, as the required minimum feature size for the patch is sub-100 μm , it is not possible to manufacture these beamsplitters for upper-millimeter-wave frequencies (*ca.* 100 to 300 GHz) using currently available desktop 3-D printers. Also, due to their

sub-wavelength periodicity, they exhibit frequency dispersion and, therefore, narrow bandwidth. Another low-cost approach for achieving free-space power splitting, at THz frequencies, employs an electrically thin-film metal-coated dielectric substrate. For example, a sub-skin depth layer of silver paint can be deposited onto a low-density polyethylene substrate [11].

The holey beamsplitter design is inspired by the concept of the polka dot beamsplitter, by exploiting Babinet's principle to create the complementary structure (i.e., the metal patches are replaced by a metal mesh), which avoids the need for printing the non-metalized areas of the substrate. This solution allows the beamsplitter to be manufactured using conventional FDM 3-D printing and metal plated using gold leaf gilding techniques [2].

FDM 3-D printing has poor dimensional precision and print consistency for sub-millimeter minimum feature sizes. Moreover, a single layer of gold leaf (having a manufacturer's quoted thickness of $0.12\text{ }\mu\text{m}$) exhibits less than one skin depth across G-band (ranging from 0.167 to $0.210\text{ }\mu\text{m}$); simulations for a single layer of gold leaf show that this is sufficient to provide a screening effectiveness [12] of approximately 59 dB and return loss of 0.01 dB, across G-band. A conventional approach for achieving equal transmittance and reflectance was not possible with a 50% metallized coverage area. Due to lack of computational resources to undertake the necessary 3-D electromagnetic modelling, an iterative 'trial and error' design-manufacture-test cycle was adopted; fortunately, this solution lends itself to rapid prototyping using 3-D printing. The key dimensions and photographs of the final version of the beamsplitter are shown in Figure 6.2.

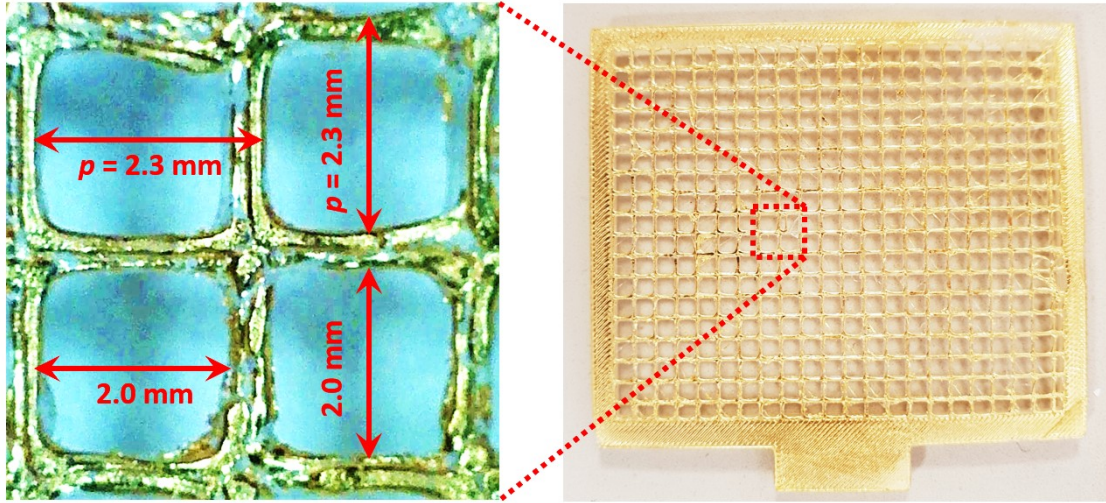


Figure 6.2: 3-D printed and gold-leaf gilded holey beamsplitter: (left) zoomed-in photograph; and (right) photograph of the complete component.

It is important to note that the beamsplitter's lattice constant $p = 2.3 \text{ mm} \gg \lambda_{\text{max}}(140 \text{ GHz})/10 = 0.2 \text{ mm}$, which means it cannot be treated as a homogeneous subwavelength metasurface (which would otherwise require $p < \lambda/10$); as such, with $p \sim \lambda$, the beamsplitter design falls under the heading of quasi-optics (which excludes both simple electromagnetic theory and simple geometrical optics ray tracing).

The power transmittance ($\mathcal{T} = |S_{21}|^2$) and reflectance ($\mathcal{R} = |S_{31}|^2$) measurement setup is shown in Figure 6.3. An ideal, lossless beamsplitter has $\mathcal{T} = \mathcal{R} = 0.5$, while the primary objective is just to achieve $\mathcal{T} = \mathcal{R}$ across G-band. Absorptance ($\mathcal{A}$), which includes all losses (e.g., scattering, diffraction, ohmic and non-ideal beam collimation), is represented by:

$$\mathcal{A} = 1 - |S_{11}|^2 - |S_{21}|^2 - |S_{31}|^2 \approx 1 - |S_{21}|^2 - |S_{31}|^2 \quad (6.2)$$

With 2-port measurements, the input return losses are $-10 \log |S_{11}|^2 < 20 \text{ dB}$ across more than 80% of G-band and so the simplifying approximation in (2) can be justified.

6.1.3 Test Platforms

To test the 3-port holey beamsplitter, at its intended oblique incidence angle of 45° , two 2-port measurements are needed; with a common input port (Port 1) and two output ports (Port 2 and Port 3), as illustrated in Figure 6.3. Both output ports need to face the input port, due to the limited cable lengths between the VNA and frequency extension heads, and also to avoid unnecessary movement of the cables [2].

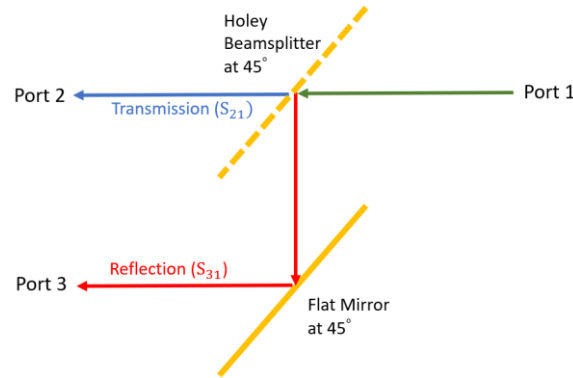


Figure 6.3: Illustration of beam paths for testing the holey beamsplitter.

To accommodate these requirements, a bespoke reconfigurable quasi-optical 3-port test platform is required. The 3-D printed plug and play test platform has pre-aligned slots for the horn antennas (in their 3-D printed adapters) and collimating plano-convex lenses. The beams transmitted through and reflected by the beamsplitter are focused by plano-convex lens into the corresponding commercial horn antennas at Port 2 and Port 3, respectively; with the latter, a flat mirror is needed to redirect the beam and this should be positioned parallel to the beamsplitter. The 3-D printed 3-port test platform is shown in Figure 6.4, without a device under test.

The input and internal beams are all vertically polarized. To evaluate the beamsplitter's polarization response, a bespoke quasi-optical rotating test platform is 3-D printed and fitted with a rotational positioner (having 3-D printed alignment markers at every 5° intervals of rotation, over a 90° arc). Here, the intended incidence angle is fixed at 0° . The 3-D printed quasi-optical rotating test platform is shown in Figure 6.5.

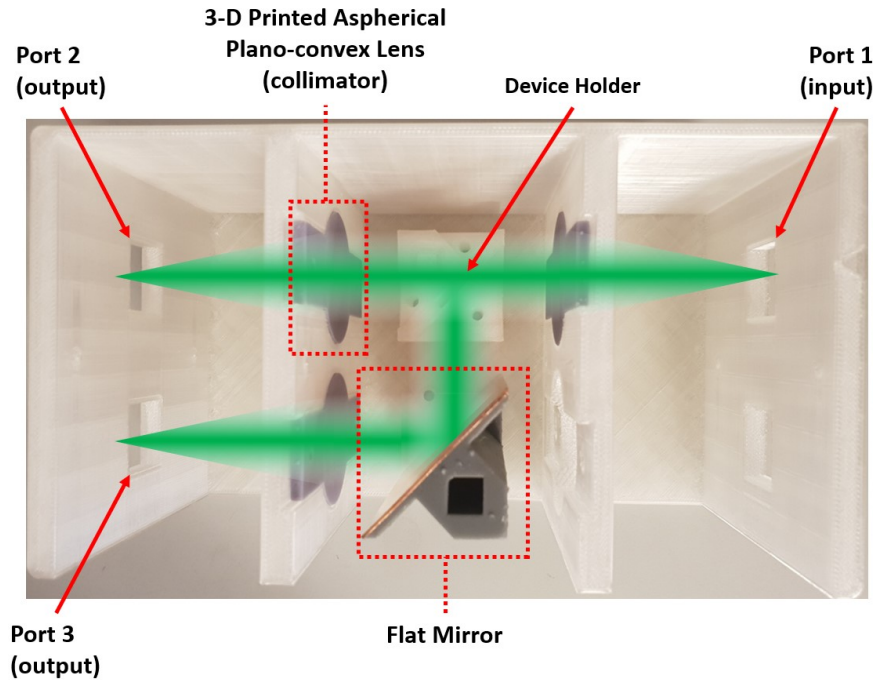


Figure 6.4: 3-D printed reconfigurable quasi-optical 3-port test platform.

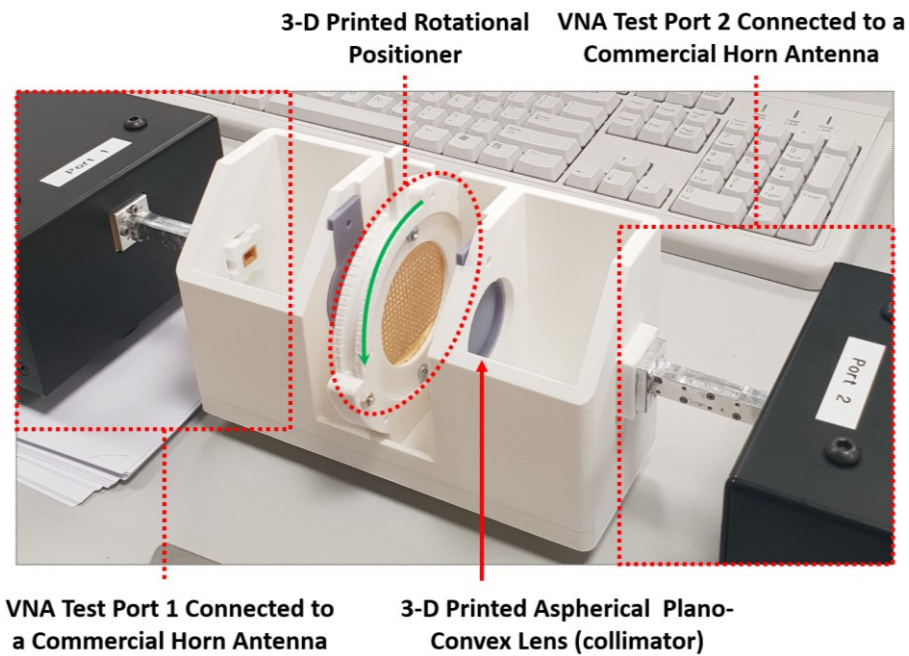


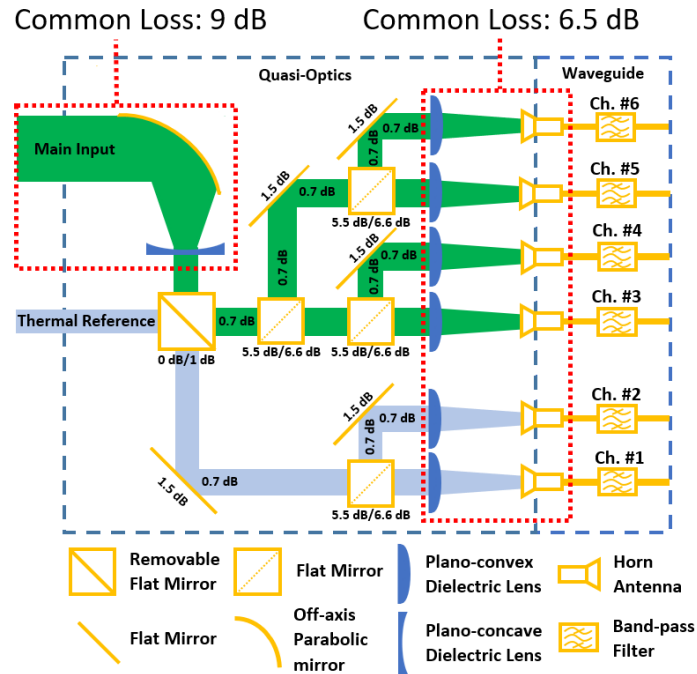
Figure 6.5: 3-D printed quasi-optical rotating test platform with holey beamsplitter as the device under test.

6.1.4 Subsystem Housing

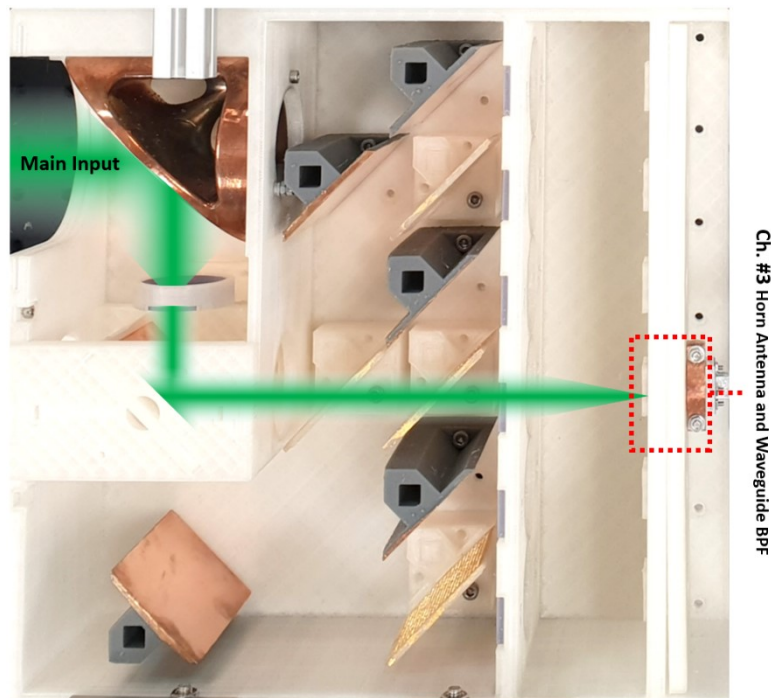
The complete single-piece subsystem housing is designed to have external dimensions of 279 mm $\times$ 265 mm $\times$ 92 mm, integrating 30 individual components. The prototype architecture is designed to support multiple iterations on the development of each component. For this reason, the plug and play methodology employ pre-aligned snap-in brackets for each component.

A block diagram for the compact multi-channel subsystem architecture is illustrated in Figure 6.6(a), with associated beam paths for each of the six channels. In addition to the main input, an auxiliary window is included with the option for the subsystem channel receivers to ‘see’ external hot or cold reference sources. The removable flat mirror has two settings. When inserted, Channels #1 and #2 can ‘see’ the thermal reference sources (blue beams in Figure 6.6(a)), while Channels #3 to #6 ‘see’ the main input (green beams in Figure 6.6(a)). In contrast, when removed, Channels #1 and #2 ‘see’ the main input, while Channels #3 to #6 ‘see’ the thermal reference sources. Although the removable flat mirror settings were changed manually (for convenience), the 3-D printed housing accommodates an electrically-actuated switching mechanism; this could then be used to implement a Dicke switch for radiometry applications.

The fully 3-D printed quasi-optical housing with integrated components for the proof-of-concept demonstration is shown in Figure 6.6(b). When compared to using a commercially-machined breadboard and off-the-shelf optical components and their positioners, 3-D printing for prototyping of a complete front-end subsystem offers the important advantage of providing a much more compact solution. In addition, component assembly and alignment are effortless as discussed in Chapter 5. It can be seen in Figure 6.6(b) that a near-optimal configuration is achieved in terms of size and compactness.



(a)



(b)

Figure 6.6: 3-D printed quasi-optical housing fitted with all 3-D printed components: (a) block diagram; and (b) photograph (plan view) with beam illustration for Ch. #3 measurement.

6.2 Fabrication

Table 6.2 lists the various components, from the test platforms and subsystem housings to individual components, and their methods of implementation [2]. In general, each component has unique methods of implementation; the rationale for this has already been discussed in detail (Chapter 5).

Table 6.2 : Component 3-D printing and metallization implementation

Component	CAD Drawing	Slicer	Printer Technology	Printer Model	Metallization
Housings [2]	Autodesk Fusion 360 [13]	IdeaMaker [14]	FDM	Raise Pro 2 [15]	-
OAPM [2]		PreForm [16]	SLA	Form 3 [17]	Copper Commercial Electroplating
Flat Mirrors					Copper Foil Lamination
Lens [2]		Chitubox [18]	MSLA	Elegoo Mars 2 Pro [19]	-
Antennas [2]					Copper Commercial Electroplating
Filters					
Beamsplitter	OpenSCAD [20]	Simplify 3D [21]	FDM	Prusa MK3 [22]	Gold Leaf Gilding

Details will now be given for the new components not previously reported. With the flat mirrors, to avoid the relatively high cost of commercial electroplating, a single sheet of $\sim 25\text{ }\mu\text{m}$ thick copper foil is simply laminated onto the flat front-facing surface of the mirror. Since the band-pass filters have small and complex internal features, commercial electroplating of $\sim 50\text{ }\mu\text{m}$ thick copper is chosen [2].

Finally, as the holey beamsplitter requires dimensional adjustments with its iterative cycle, the script-based CAD drawing software, OpenSCAD [20], is used. Being only $500\text{ }\mu\text{m}$ thick, this component is too delicate for commercial electroplating; hence, gold leaf gilding is employed [2]. Applying more than one layer of gold leaf causes glue to clump between the holes, causing uneven surfaces. Therefore, only a single layer of gold leaf is gilded onto one side of the holey beamsplitter's 3-D printed skeletal frame.

6.3 Component Measurement

All electromagnetic measurements are undertaken at the U.K.'s National Physical Laboratory (NPL), using their Keysight Technologies PNA-X N5247B vector network analyzer (VNA) – with VDI WR-5.1 frequency extension heads [23].

Two different calibration methodologies are needed: Thru-Reflect-Line (TRL) calibration in conventional metal-pipe rectangular waveguide, with reference planes at the VNA frequency extension heads; and Thru-Reflect-Match (TRM) calibration in free space having two reference plane options.

After completing the TRL calibration, commercial Flann Microwave Ltd 20 dBi standard gain rectangular horn antennas are fixed to each port [24]. The antennas, with their interface adapters, are then slotted into the quasi-optical test platform shown in Figure 6.4. This set-up still has its calibration reference planes located at the flanges of the frequency extension heads. Therefore, any subsequent measurements will include the adverse effects due to the non-ideal horn antennas, lenses and free-space channel, illustrated in Figure 6.7.



Figure 6.7: Illustration of the limitation for TRL calibration at the VNA's frequency extension heads for the free-space quasi-optical measurement platform.

To avoid these adverse effects, while isolating the device under test in free-space, TRM calibrations are employed for some of the measurements, which are widely used for free-space material characterization [25]. TRM calibration does not require movement of any components during its calibration process [26], making it ideal for the fixed position test platform.

As shown in Figure 6.3, the beamsplitter represents a 3-Port network that requires both transmittance and reflectance measurements. However, since the VNA and its frequency extension heads only support 2-Port testing, two different sets of measurements are required, each having its own calibration. For the transmittance measurements, TRM Calibration #1 is adopted, as illustrated on the left-hand side of Figure 6.8. For the reflectance measurements, TRM Calibration #2 is adopted, as illustrated on the right-hand side of Figure 6.8. For both calibrations, 3-D printed flat mirrors and commercial pyramidal absorbers (TK Instruments [27]) is used for the Reflect and Match standards, respectively. These calibrations provide an initial common reference plane positioned midway between the lenses.

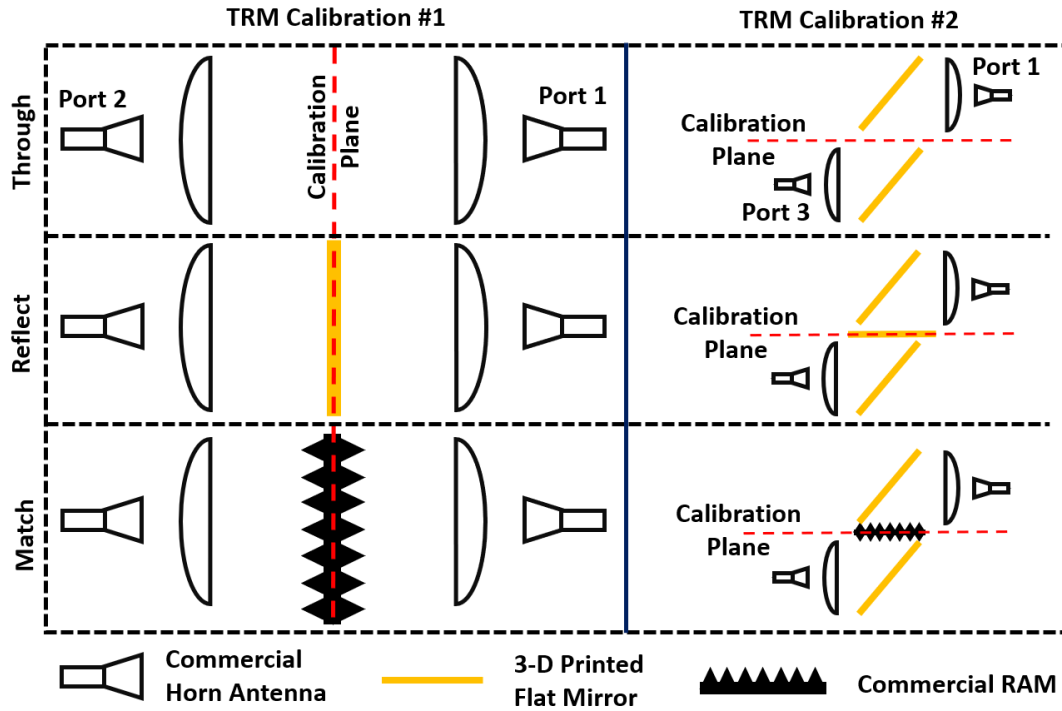


Figure 6.8: Illustration of free-space TRM Calibration #1 (left) and #2 (right) for the reconfigurable quasi-optical measurement platform.

6.3.1 Flat Mirrors

Using the quasi-optical test platform shown in Figure 6.4, with TRM Calibration #1 shown in Figure 6.8, the performance of the flat mirror is evaluated. A pair of identical flat mirrors, at an oblique incidence angle of 45° , creates a thru-line, as illustrated by the insert in Figure 6.9. The initial TRM Calibration #1 reference plane is effectively split in two and shifted from the center of the mirrors to their outer edges. As a result, the transmission measurement, in Figure 6.9, represents the combined losses from both flat mirrors (e.g., misalignment, ohmic and surface defect-related diffuse reflection scattering) and non-ideal beam collimation exacerbated by excess path lengths between mirrors. It can be seen that a single flat mirror is expected to have an associated loss of 1-2 dB. However, it is important to note that this result only represents a rough pseudo-qualitative estimation of the flat mirror's performance.

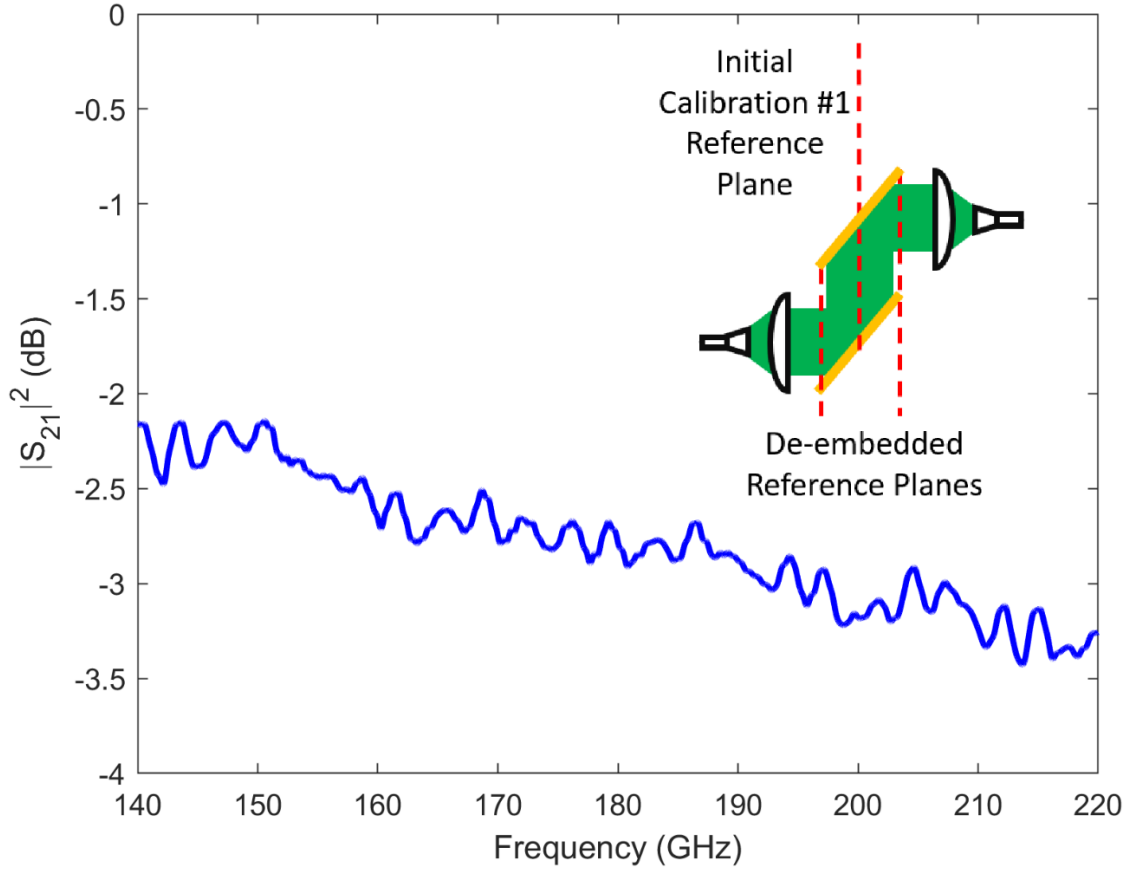
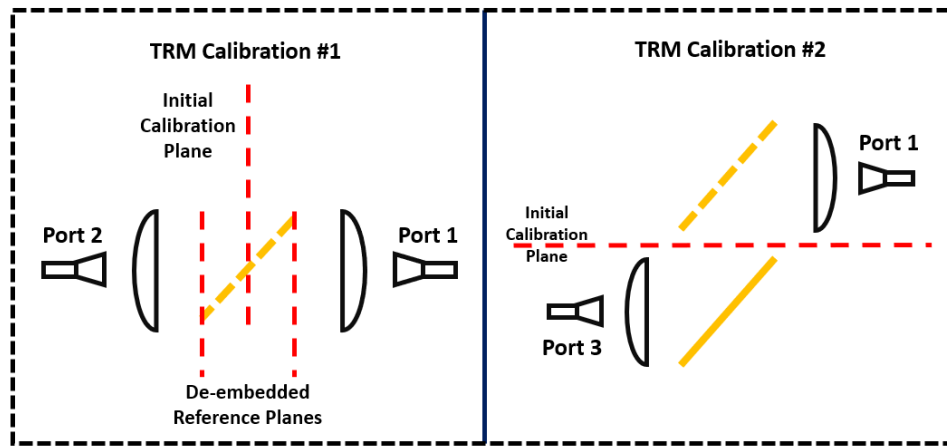


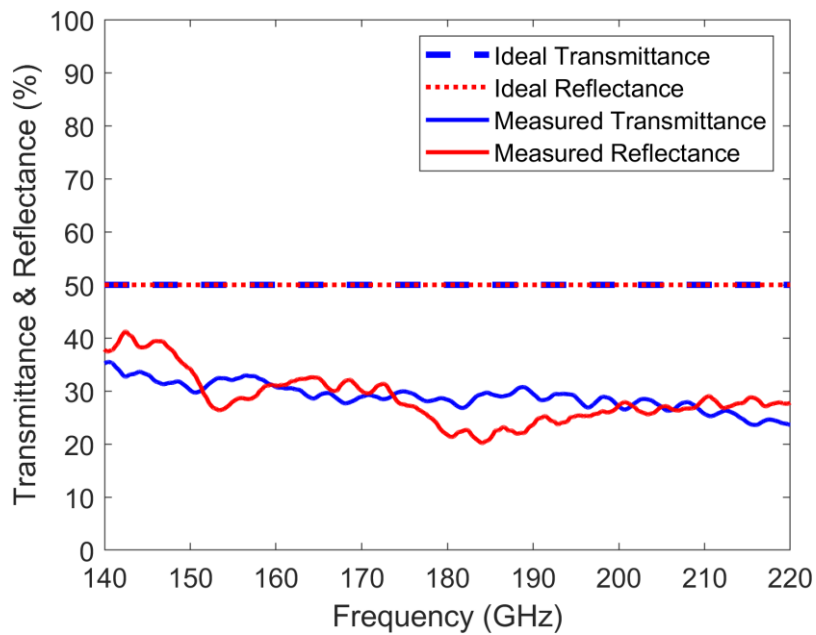
Figure 6.9: Measured transmission through two flat mirrors at 45° incidence using TRM Calibration #1 with additional de-embedding of the reference planes (running averaging applied across 3.2 GHz).

6.3.2 Holey Beamsplitter

Using the same 3-port quasi-optical test platform, shown in Figure 6.4, the 3-D printed holey beamsplitter is now evaluated. The measurement set-up employs both TRM Calibrations #1 and #2, illustrated in Figure 6.10(a), for the transmittance and reflectance measurements, respectively, with results shown in Figure 6.10(b). With the former, the initial TRM Calibration #1 reference plane is effectively split and shifted from the center of the beamsplitter to its outer edges.



(a)



(b)

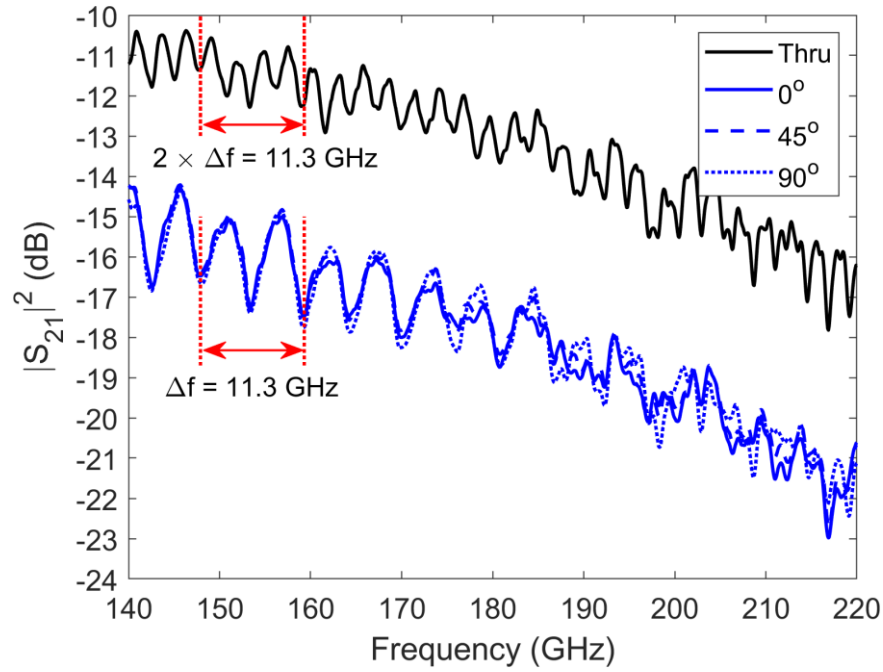
Figure 6.10: Holey beamsplitter measurements: (a) illustration of the test set-ups using TRM Calibration #1 (with additional de-embedding of the reference planes) and #2 for measuring transmittance and reflectance, respectively; and (b) measured transmittance and reflectance at 45° incidence (running averaging applied over 3.2 GHz).

Figure 6.10 shows that the holey beamsplitter is able to equally divide the incident power; fulfilling the primary objective. The maximum difference between transmittance and reflectance is less than 9%, across G-band. Ideally, absorptance should be 0%. In practice, due to a number of factors (e.g., diffraction, diffuse scattering, non-ideal beam collimation and ohmic losses), the holey beamsplitter has approximately 30-50% of absorptance. Table 6.3: Beamsplitter performance comparison shows the performance of ideal, typical commercial optical and the holey beamsplitters. When compared to the commercial optical polka dot beamsplitter, the millimeter-wave proof-of-principle demonstrator has approximately 20-30% more absorptance.

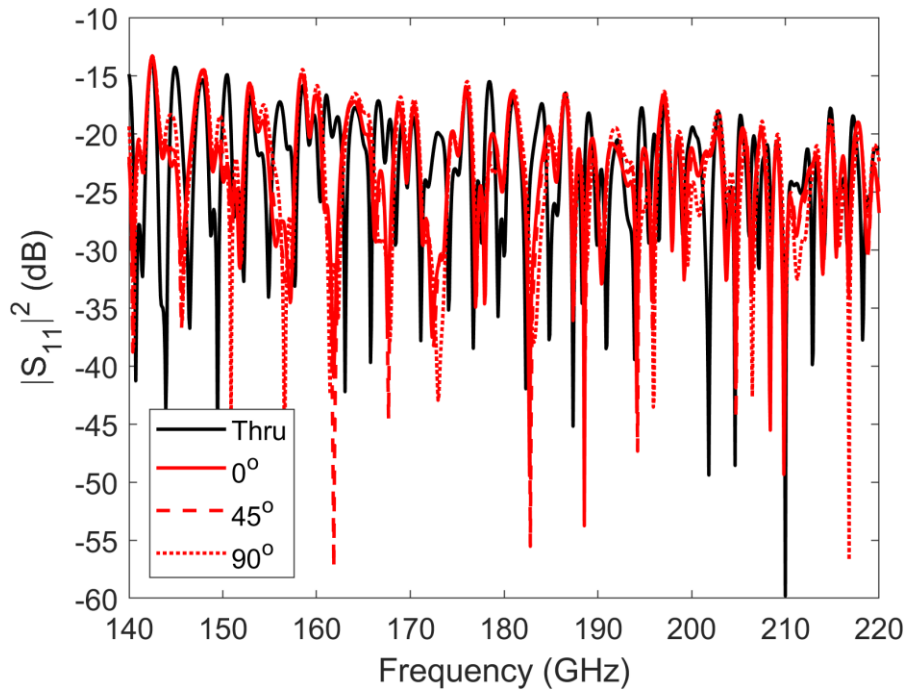
Table 6.3: Beamsplitter performance comparison

Response (%)	Ideal	Commercial Polka Dot (Optical) [4]	3-D Printed Holey (G-band)
Transmittance, $\mathcal{T}$	50	40 – 50	25 – 35
Reflectance, $\mathcal{R}$	50	40 – 50	25 – 40
Absorptance, $\mathcal{A}$	0	10 – 20	30 – 50
$ \mathcal{T} - \mathcal{R} $	0	5 – 10	5 – 10

With the vertically-polarized collimated beams, the S-parameters are measured at different transverse angles of rotation, using the 3-D printed polarization test platform, shown in Figure 6.5. For this measurement, conventional TRL calibration at the VNA's frequency extension heads is performed. As shown in Figure 6.11(a), the beamsplitter has identical transmission responses at three transverse angles of rotation (0° , 45° and 90°); confirming polarization independence.



(a)



(b)

Figure 6.11: Holey beamsplitter transmission measurements for different polarizations (transverse angles of rotation) with TRL calibration: (a) transmission; and (b) reflection (no running averaging applied).

The normalized transmittance of the beamsplitter at normal incidence, with quasi-optical test fixture, $\mathcal{T}_n(\text{dB}) \cong |S_{21}|^2_{\text{thru}}(\text{dB}) - |S_{21}|^2(\text{dB})$, where $|S_{21}|^2_{\text{thru}}(\text{dB})$ corresponds to the transmittance when the beamsplitter is removed. Here, $\mathcal{T}_n(\text{dB}) \sim 5 \text{ dB}$ at 180 GHz, for all three rotation settings; this being comparable to the measured transmittance $\mathcal{T} = 5.5 \text{ dB}$ at 180 GHz, extracted from Figure 6.10, at the oblique incidence angle of 45° . As seen in Figure 6.11b), the reflection responses at all three transverse angles of rotation are also almost identical. It is important to note that, due to the high insertion loss for each lens ($>5 \text{ dB}$, discussed in the next sub-section), there is almost no measured difference in reflectance seen in Figure 6.11(b), with or without the beamsplitter.

Note that the observed in-band Fabry-Pérot ripples, having frequency period Δf , correspond to standing-wave reflections between two spatial boundaries that are separated by a theoretical free-space distance $\Delta l = c/\Delta f$. Without the beamsplitter, the extracted value of $\Delta l \cong 53 \text{ mm}$ corresponds to the spatial distance of 55.0 mm between the flat sides of both plano-convex lenses. With the beamsplitter inserted, the extracted value of $\Delta l \cong 26.5 \text{ mm}$ corresponds to the spatial distance of 27.8 mm between the beamsplitter and the flat side of the plano-convex lens. This shows that the metalized side of the beamsplitter is positioned exactly midway between the two plano-convex lenses.

6.3.3 Plano-Convex Lenses

Using the 3-D printed test platform shown in Figure 6.4, with conventional TRL calibration at the VNA's frequency extension heads, the performance of a lens pair is evaluated. Similar to the arrangement shown in Figure 6.7, transmission between two commercial horn antennas, with and without the lenses, is measured. From Figure 6.12, it can be seen that adding the lenses provides between 6 to 17 dB of 'gain enhancement'. This clearly shows that the lens is performing as expected. The commercial antenna and 3-D printed lens pair exhibits, in situ, approximately 5-8 dB of insertion loss across G-band. Note that the extracted value of $\Delta l \cong 62.5 \text{ mm}$ corresponds to the spatial distance of 65.0 mm between the flat sides of both plano-convex lenses.

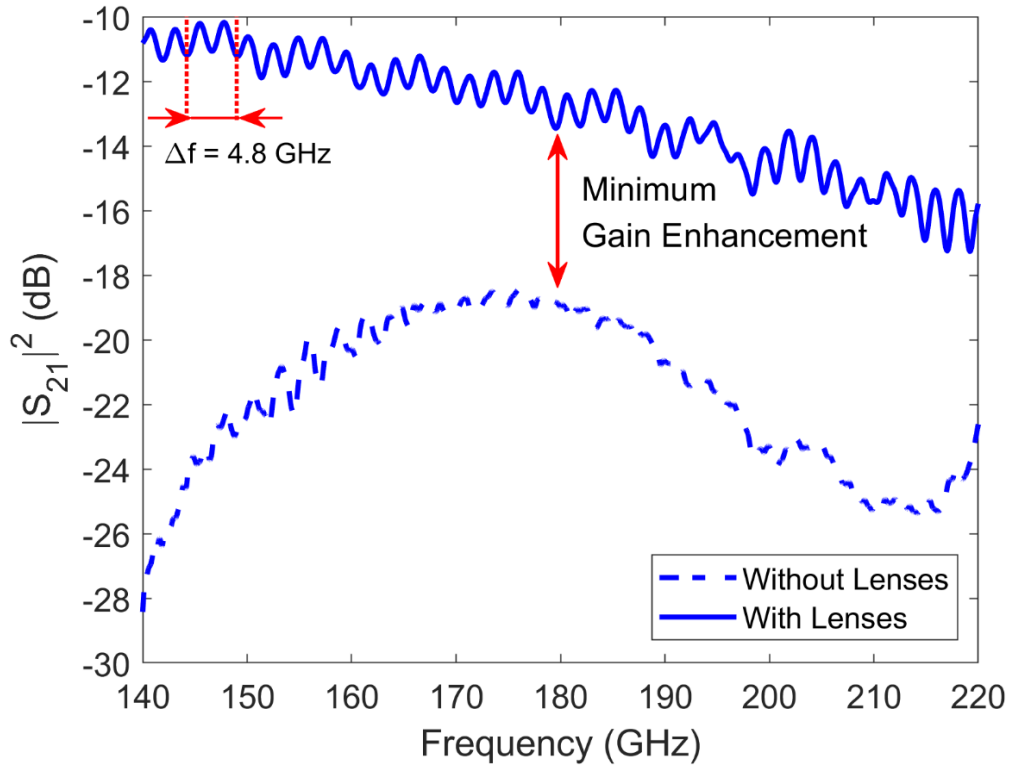
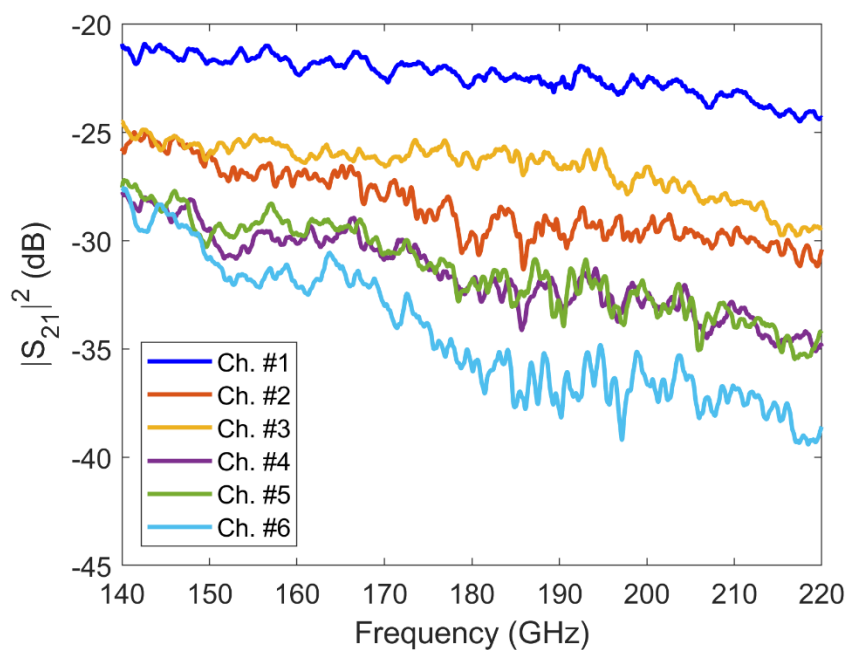
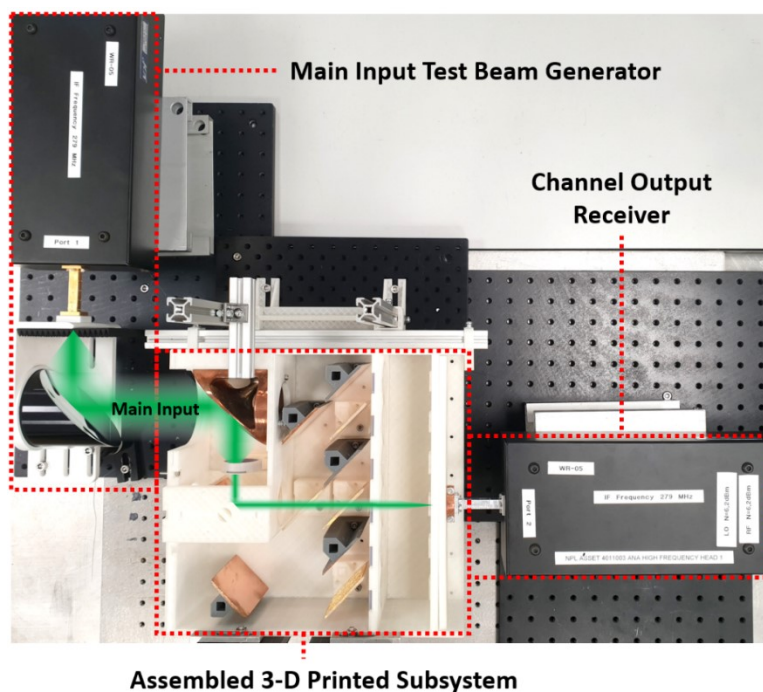


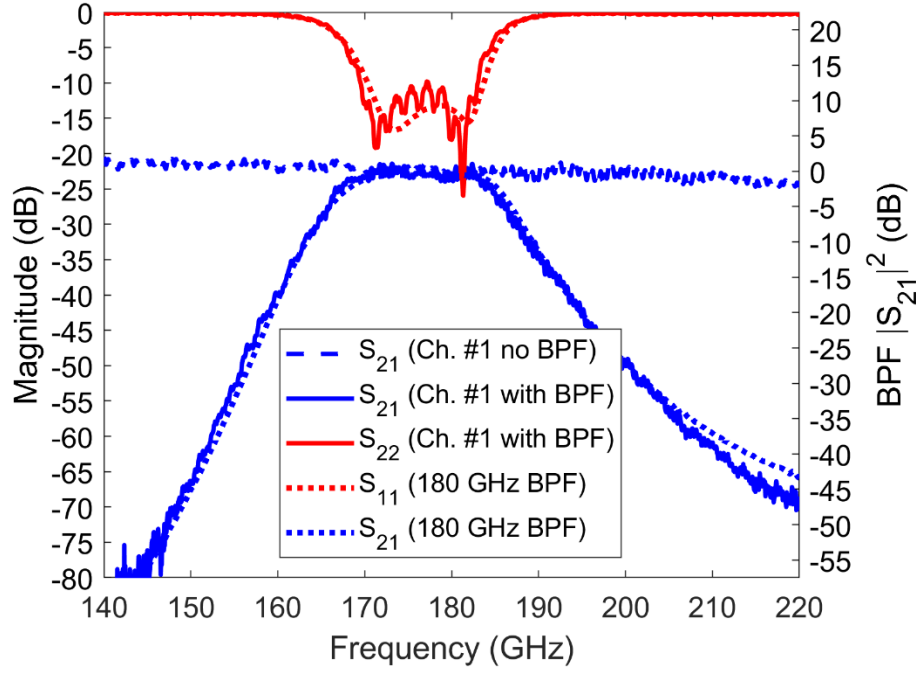
Figure 6.12: Plano-convex lens transmission measurements (running averaging applied over 3.2 GHz).

6.4 Integrated Subsystem Measurement

With conventional TRL calibration at the VNA's frequency extension heads, the performance of the fully assembled 3-D printed integrated subsystem is evaluated. At Port 1, commercial rectangular horn antenna and 90° OAPM are added to emulate the collimated main input beam [2]; adding an estimated 2 dB of associated losses. Port 2 is connected directly to one of the six output channels, to emulate the channel output receiver. The complete front-end measurement set-up and its results are shown in Figure 6.13.

The measured transmission results for the six channels, without their filters, are shown in Figure 6.13(b). The lowest transmission loss is observed with Ch. #1, as expected; between 21 and 24 dB across G-band. The highest transmission loss is observed with Ch. #6; between 27 and 40 dB across G-band. Clearly, performance degrades as frequency increases, due to non-ideal beam collimation,





(c)

Figure 6.13: Assembled subsystem measurements: (a) photograph of the quasi-optical test set-up; (b) transmission losses for each channel without filters (running averaging applied over 5 GHz); and (c) measured responses for Ch. #1 integrated with an example 180 GHz 3-D printed band-pass filter (no running averaging applied).

Due to the complexity of the integrated subsystem, it is nearly impossible accurately quantify the loss contributions associated with each and every component. However, using the measurements in Figure 6.13(b) and with information for individual components (from Section 6.3), it is possible to make a pseudo-quantitative loss budget analysis for the complete subsystem.

An estimated loss for each component type is assigned and their total loss is then compared with that measured for each channel. From this, it can be speculated that both the flat mirrors and beamsplitters create an additional loss of ~ 0.7 dB at 180 GHz in their reflected paths, because the beams are not perfectly collimated. This is further compounded by both external and internal misalignments of the quasi-optical components. The results of the preliminary analysis are shown in Table 6.4: Pseudo-quantitative loss budget analysis at 180

GHz for each element and the complete assembled subsystem (without BPFs). and Figure 6.6(a). With the exception of Ch. #2, it can be seen that there is good agreement between the estimated and measured total losses.

Table 6.4: Pseudo-quantitative loss budget analysis at 180 GHz for each element and the complete assembled subsystem (without BPFs).

elements (Cumulative)	Estimated Losses (dB)					
	Ch. #1	Ch. #2	Ch. #3	Ch. #4	Ch. #5	Ch. #6
Common Input Stage	9	9	9	9	9	9
Removable Flat Mirror	-	-	1	1	1	1
Flat Mirrors	1.5	3	-	1.5	1.5	3
Beamsplitter Transmissions	5.5	-	11	5.5	5.5	-
Beamsplitter Reflections	-	6.6	-	6.6	6.6	13.2
Common Output Stage	6.5	6.5	6.5	6.5	6.5	6.5
Non-ideal Collimation	0.7	2.1	0.7	2.1	2.1	3.5
Estimated Total Loss	23	27	28	32	32	36
Measured Total Loss	23	30	27	32	32	36

6.5 Discussion and Conclusion

In this chapter, component integration concept was demonstrated by employing a complete plug and play 6-channel subsystem for operation between 140 to 220 GHz. This sub-THz front-end architecture has successfully integrated 30 individual components: 1× OAPM, 1× plano-concave lens, 1× reconfigurable flat mirror, 5× flat mirrors, 4× beamsplitters, 6× plano-convex lenses, 6× horn antennas and 6× BPFs. It was found that a blend of different 3-D printing

technologies and plating methods were needed, depending on the type of component.

In this chapter, bespoke quasi-optical TRM calibrations and measurement schemes were introduced with the 3-D printed test platforms. The 3-D printed flat mirrors demonstrated total losses between 1-2 dB across G-band. The holey beamsplitter was able to divide the input beam into two output beams with equal power. A pair of 3-D printed plano-convex lenses gave up to 17 dB of gain enhancement across G-band. Four BPFs showed good transmission and reflection measured responses. Finally, two test platforms and integrated subsystem housings were all 3-D printed as single pieces, to support plug and play development, while providing sufficient beam alignment accuracy.

Having been able to establish a pseudo-quantitative loss budget for the individual elements, future work on minimizing associated losses can now be investigated. For example, measurements of Ch. #1, having a 180 GHz BPF included, demonstrate a total measured insertion loss of 23 dB for the first proof-of-concept prototype. While this is likely to be too large for most applications, the link budget helps to identify where significant improvements can be made.

The lenses were made from a relatively lossy 3-D printable dielectric material and so the plano-convex lens exhibited a transmission loss that is >5 dB. Therefore, a 3-D printable dielectric that exhibits a much lower loss tangent will reduce transmission losses; while a greater refractive index will reduce the absorptance due to having a thinner lens, but the impedance-mismatch reflection losses will dominate and increase the net transmission loss. Alternatively, focusing mirrors (e.g., OAPMs) can completely replace the lenses, but this will dramatically increase the size and mass of the assembled subsystem; these and other engineering compromises can be explored in future studies.

Another component worthy of further investigation is the beamsplitter. After considerable investigation, there does not appear to be an obvious 3-D printing alternative to the solution. Nevertheless, with the holey beamsplitter approach, optimization of the internal macroscopic geometrical features (using full-wave electromagnetic simulation software), better material selection and

advanced manufacturing techniques, will inevitably reduce its loss characteristics.

One of the major contributors to loss is non-ideal beam collimation and misalignment. Replacing the rectangular horn antenna with a 3-D printed corrugated conical horn antenna will help to preserve the bi-axial symmetry of the beam [2]. Moreover, the manufacturing tolerances associated with the FDM printed housing, plug and play interfaces and adapters can be improved with more advanced, commercial-grade 3-D printing technologies.

In conclusion, when compared to using a commercially machined breadboard and off-the-shelf optical components and their positioners, 3-D printing for prototyping of a complete front-end subsystem offers the important advantage of providing a much more compact solution. In addition, component assembly and alignment are effortless.

This work opens-up new opportunities for rapid prototyping and small-batch production of complete low-cost quasi-optical upper-millimeter-wave subsystems. Indeed, with the continual advancements in all aspects of affordable 3-D printers, and their materials, better resolution and higher performance can be expected in the near future.

6.6 Acknowledgements

Work presented in this chapter was submitted on the paper: “3-D printed plug and play prototyping for low-cost sub-THz subsystem” by **S.-H. Shin**, R. Payabulli, L. Zhu, M. Stanley, X. Shang, N. M. Ridler, and S. Lucyszyn, (*IEEE Access, Accepted for Publication*)[1]. The research on this work has been performed under the supervision of Prof. S. Lucyszyn and he went over every word of the original article. The BPF used for final subsystem measurement was designed by R. Payabulli. Measurements at NPL were performed under the supervision and help of M. Stanley, X. Shang, and N. M. Ridler.

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7 Bespoke Mm-wave Single Pixel Camera

7.1 Introduction

In this chapter, an mm-wave raster scanning single-pixel camera will be custom-built. This work has been published in an article by Shin *et al.* [1]. It will be used to benchmark a commercial THz camera which is one of the main tools used for characterizing various 3-D printed quasi-optical components such as lens and mirror. These 3-D printed mm-wave components involve iterative processes of the design-manufacture-test cycle. Therefore, rather than using an outsourced full-size anechoic chamber every time, an in-house mm-wave camera can be used as a preliminary measurement tool. The bespoke mm-wave camera is currently being used for some collaborative works on developing 3-D printed mm-wave antennas. It is also planned to be used for future studies on characterizing or finding 3-D printing defects such as air bubbles and layer delamination on various quasi-optical components such as zone-plates, graded index lenses and mirrors.

Radiometric imaging technologies are based on measuring thermal noise, either naturally emitted by a body (using a ‘passive radiometer’) or enhanced by illumination of the body using external sources, whereby the coherent signal energy is absorbed by the body and re-radiated as additional incoherent thermal noise energy (using an ‘active radiometer’). Radiometry in the thermal infrared (IR) spectrum (for example, at frequencies from 20 THz up to 40 THz, corresponding to wavelengths in free-space from 14 μm down to 7 μm) has the advantage of relatively good image resolution and fidelity with commercially available cameras, but has limitations on penetration depth.

The sub-THz band (i.e., any frequencies below 1 THz) offers the advantage of unprecedented penetration depth, for seeing through materials such as fabric, ceramic, plastic, leather and paper/cardboard. Although, above *ca.* 600 GHz, electromagnetic scattering off woven fabrics is still a major limitation. In addition, as a general rule of thumb, the cost of implementing technologies increases with increasing frequency, up into the infrared spectrum. For this reason, moving down the frequency spectrum gives improved penetration depth

and can potentially be cheaper, but with low spatial resolution from a compact camera. The millimeter-wave spectrum (30 GHz up to 300 GHz, corresponding to wavelengths in free-space from 10 mm down to 1 mm) offers a compromise in terms of penetration depth against spatial resolution; technologies at these frequencies will be the focus of this work.

Operating at millimeter frequencies has a number of other important issues for their practical implementation. The first is that Planck's law shows that the thermal 'brightness' of a body diminishes as frequency decreases; having its peak value within the IR spectrum for a body at room temperature. In contrast, atmospheric attenuation tends to improve as frequency decreases; although there are spectral peaks that should be avoided, due to molecular resonances (oxygen and water absorption bands) – this explains why millimeter-wave systems operate below *ca.* 40 GHz and between 70 GHz and 100 GHz (avoiding the broad and strong oxygen absorption band, centered at approximately 60 GHz) [2].

There are a number of room-temperature radiometric imaging systems available on the open market, used mainly for industrial non-destructive testing and security screening applications – based on the following generic technologies: (i) very high performance, extremely high-cost heterodyne radiometers; (ii) medium performance, very high cost pre-amplified direct detection radiometers; and (iii) low performance, medium cost non-pre-amplified direct detection radiometers, which requires additional critically placed 'active' source(s) of illumination. An example of (iii) is the US-Russian TeraSense camera; the technical details of which are inferred from their recent publications [3]-[5]. Their mid-range Tera-1024 (sub-)THz camera has a 2-D focal plane array (FPA) with 32 x 32 pixels. Each pixel has a plasmonic detector, made from a patented GaAs/AlGaAs heterostructure technology [3]. The embedded two-dimensional electron system (2DES) channel (having a resistance of 100 Ω [4]) is grown on a semi-insulating GaAs substrate. Two inner gates (with ~ 0.5 V of asymmetric DC bias) create a photon microcavity, having a width and length of 15 μm and 20 μm , respectively. These input electrodes are fed by a broadband on-chip log-periodic antenna, having an external diameter of 1.5 mm [4]. With a 1.5 mm pitch between pixels, the half-wavelength spacing provides optimal

performance at 100 GHz. The microcavity light modes couple with the 2DES plasma modes, having a sub-half-micron vertical separation distance, to excite confined relativistic plasmons (damped plasma waves) that can propagate along the 2DES channel at room temperature [3]-[5]. The rectified output photo-response signal is taken from two outer (source and drain) electrodes. The amplitude of this signal provides information about the power intensity of the incident electromagnetic radiation (recorded in arbitrary units).

According to Nyquist, the theoretical minimum spatial resolution of the camera is half the wavelength of the electromagnetic radiation being observed. The TeraSense camera has a certified spectral range from 50 GHz up to 700 GHz [6]. Moreover, at a room temperature of approximately 296 K (23°C), the detector has a post-amplification voltage responsivity of 50 kV/W and noise equivalent power (NEP) of 1 nW/ $\sqrt{\text{Hz}}$ [4] from 10 GHz to 1 THz.

When compared to uncooled microbolometer (ultra-broadband thermal detector) based FPA cameras optimized for THz applications, which can require a post-detection integration time of tens of milliseconds, plasmonic detectors have a few order of magnitude shorter response times with its room temperature NEP being an order of magnitude higher, an external ‘active’ illumination source is required with both uncooled microbolometer and other commercial THz cameras.

Nevertheless, with these commercial mm-wave cameras being able to operate at millimeter-wave frequencies, this state-of-the-art technology represents a unique and valuable room-temperature imaging tool for the scientific and engineering communities; for example, in the absence of an expensive professional antenna test range [7]. It is important to independently investigate the performance of this camera within a quasi-optical environment.

7.2 Millimeter-wave Cameras

Cameras provide quick and easy ways of measuring quasi-optical features of beams, such as cross-sectional images and intensity distributions along the propagation z -axis. For example, at (sub-)THz frequencies, there are at least two

uncooled microbolometer FPA cameras on the market; both being relatively expensive. The TZCam (from i2S, France) has a 320×240 pixel array with a $50 \mu\text{m}$ pitch that provides an associated peak performance at ~ 3 THz, with a specified frequency of operation from 300 GHz to 5 THz; requiring an additional electronically-controlled mechanical shutter for zero baseline level calibration (with repeated calibration-measurement cycles needed to increase the signal-to-noise ratio) [8]. Also, the MICROXCAM-384i-THz (from INO, Canada) has a 384×288 pixel array with a $35 \mu\text{m}$ pitch that provides an associated peak performance at ~ 4 THz, with a specified frequency of operation from 94 GHz to 4.25 THz [9]. However, the next-generation of (non-pre-amplified direct detection) CMOS FPA cameras are aimed at the low-cost market. For example, the TicMOS-1kpx (from TicWave, Germany) has a 32×32 pixel array and a -3 dB bandwidth between 0.75 THz to 1 THz, with a specified frequency of operation from 100 GHz to 4 THz [10]. More information on these products is available [11]-[12].

Here, a sample commercial camera against a custom-built single-pixel camera at W-band (75 GHz to 110 GHz) is benchmarked, for image quality and aperture reflectance.

7.2.1 Single-pixel Raster Scanned

A single-pixel raster-scanning camera is custom-built, based on one millimeter-wave diode detector, 3-D scanning mechanism and an automated data acquisition module. 3-D printed brackets and mounts are employed, with an optical breadboard for precision alignment. A block diagram, illustrating each component of the single-pixel camera, is shown on Figure 7.1(a).

The camera has a square-law detector (Millitech DXP-10 Schottky barrier beam lead diode detector [13]), operating across W-band, with a standard open-ended WR-10 waveguide aperture ($2.54 \text{ mm} \times 1.27 \text{ mm}$) surrounded by radiation absorbent material (RAM).

For creating 2-D cross-sectional images, perpendicular to the propagation z -axis, the detector raster scans over the horizontal and vertical x - y plane. Figure 7.1(b) shows a simplified illustration of the raster scanning sequence.

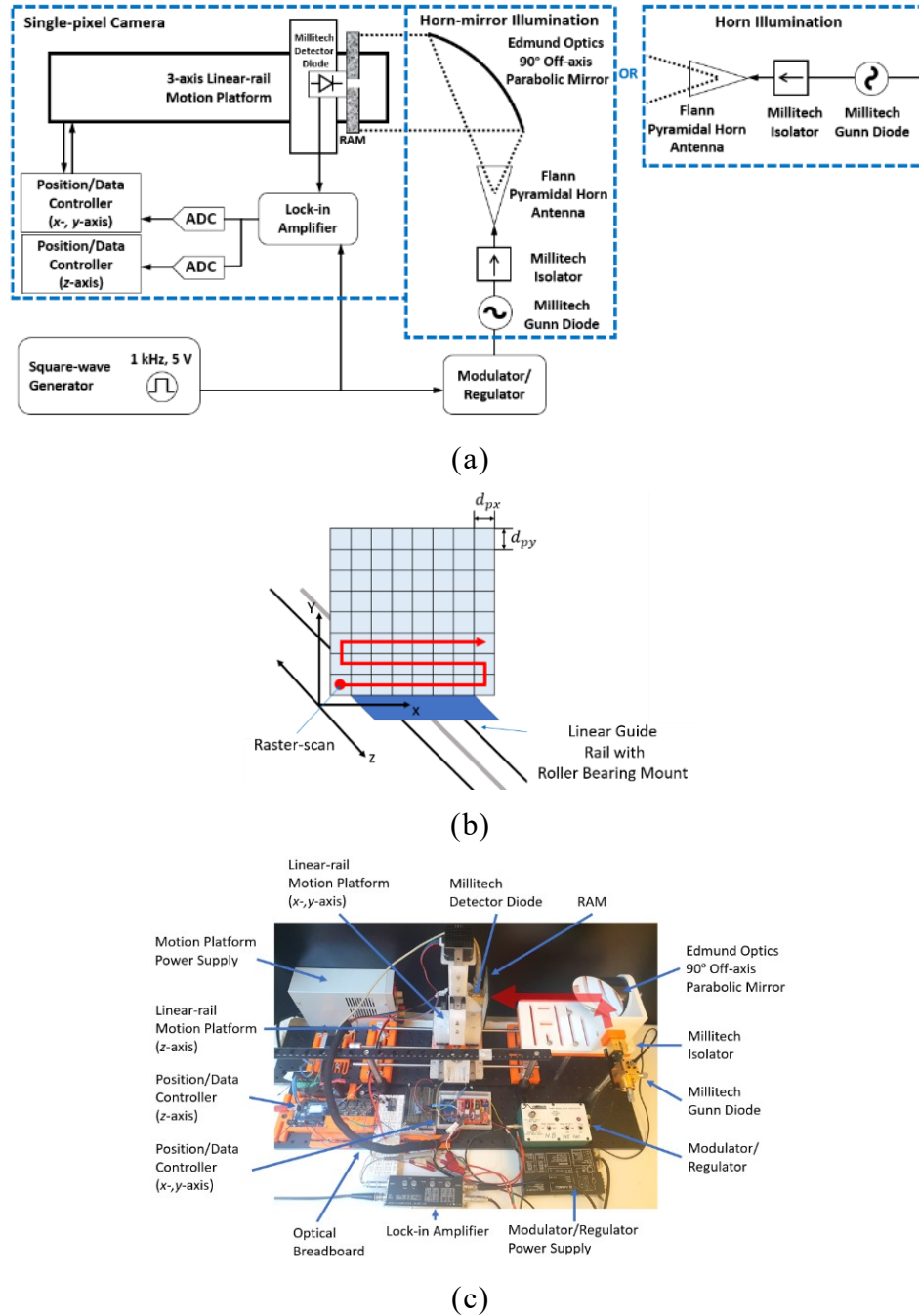


Figure 7.1 : Custom-built single-pixel raster-scanning camera: (a) basic block diagram; (b) illustration of raster scanning sequence; and (c) complete single-pixel camera system.

The 3-axes (x , y and z) scanning mechanism is actuated using three pairs of stepper motors and 8 mm lead screws. Each axis is guided with a pair of linear-rails and matching roller bearings. The stepper motors are controlled with two microcontrollers (ATmega328P and ATmega2560) [14]–[15]. Depending on the spatial resolution setting, the microcontrollers set the direction, speed and step-counts for the stepper motors. The stepper motors are synchronized to accommodate the sampling rates for the data acquisition module. The minimum achievable pixel size (d_{px} , d_{py}) with this setup is $\sim 10\text{ }\mu\text{m}$.

The basic data acquisition module employs a lock-in amplifier and two (Texas Instrument ADS1115 [16]) analog-to-digital converters (ADCs), which feed the microcontroller boards that are synchronized with the scanning mechanism. The complete single-pixel camera system is shown in Figure 7.1(c).

7.2.2 Commercial FPA

With any FPA, image quality is dependent on both spatial resolution and pixel fidelity. The latter is a function of the number and distribution of pixels that are faulty and those with large deviations in responsivity. This example commercial FPA camera has a specified permissible number of faulty pixels of up to 3% and specified pixel-to-pixel responsivity deviation from the mean value is within 20%; requiring additional image calibration/correction measures. As with the single-pixel camera, external active illumination is needed [4].

Since the camera has limited spatial resolution, faulty pixels can corrupt beam profile images. In addition, faulty pixels near the center of a beam makes it difficult to identify the location of peak intensity. Also, over a period of months, the number of faulty pixels increased. To expose and map the faulty pixels, the camera's aperture is fully illuminated by a beam at maximum exposure and 100 frames are accumulated. Initially, 24 faulty pixels were recorded, with 16 more faulty pixels identified after 30 months. The quasi-optical measurements for this work are undertaken with between 26 and 40 faulty pixels. Figure 7.2 shows the mapped locations of faulty pixels. Clearly, as the number of both dead and malfunctioning pixels increase, the need for sophisticated image processing increases and image fidelity becomes more problematic.

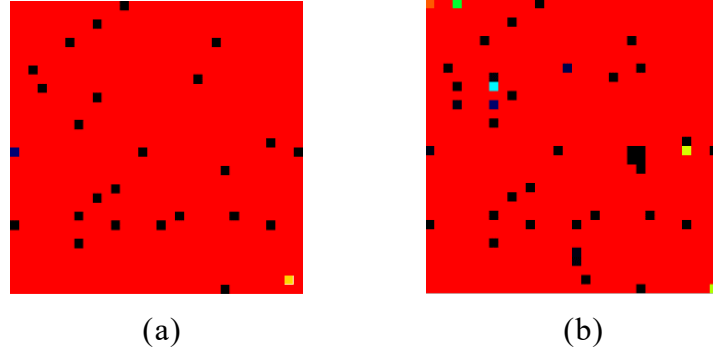


Figure 7.2 : Faulty pixel map for the camera with a count of: (a) 26, prior to this work; (b) 40, after this work. Black squares represent dead pixels (no output) and blue/green/yellow pixels give a false reading.

A performance summary for each camera is given in Table 7.1.

Table 7.1 : Performance summary for the single-pixel and typical commercial cameras

Specifications	Camera	
	Single-pixel	Commercial
Step/Pixel Size (mm)	0.2 (step)	1-2 (pixel)
Faulty Pixels	0	26
Pixel Responsibility Deviation (%)	≈ 0	≤ 20
Exposure Time	3 – 6 hrs	< 1 min
Relative Cost	low	medium

7.2.3 Beam Sources

For this study, two Gunn diode modules are used (Millitech GDM-12-WB-17M, operating at 76 GHz and Millitech GDM-10-0-17H, operating at 92 GHz [13]). In both cases, an isolator (Millitech JFL-10-NI [13]) is employed to protect the Gunn diode modules from impedance mismatch reflections. With the 76 GHz source, an additional WR-12 to WR-10 waveguide transition is inserted between the Gunn diode module and isolator. For biasing the sources, a separate Gunn Modulator Regulator (Millitech GMR) is employed. When square-wave modulation is needed, an external modulator (1 kHz, 5 V) is used to clock the GMR.

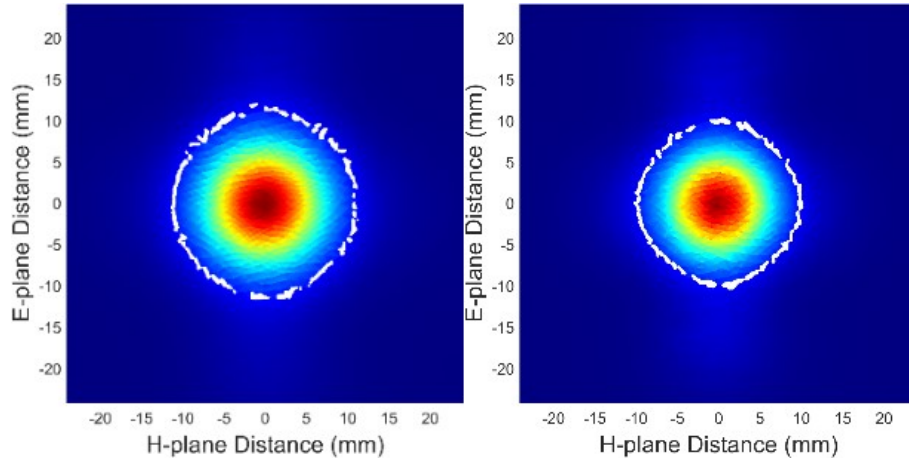
As shown in Figure 7.1(a), with a typical quasi-optical (QO) set-up, a pyramidal horn antenna (FLANN 27240-20 [17]) is employed with the Gunn diode module to radiate a diverging beam. When a collimated beam is needed, a 90° off-axis parabolic mirror (Edmund Optics 36-601 [18]) is placed 76.2 mm away from the pyramidal horn antenna's effective focal point. It is worth noting that perfect collimation is unachievable with a pyramidal horn antenna because it does not represent an ideal point source as discussed in Chapter 5.

7.3 Measuring Image quality

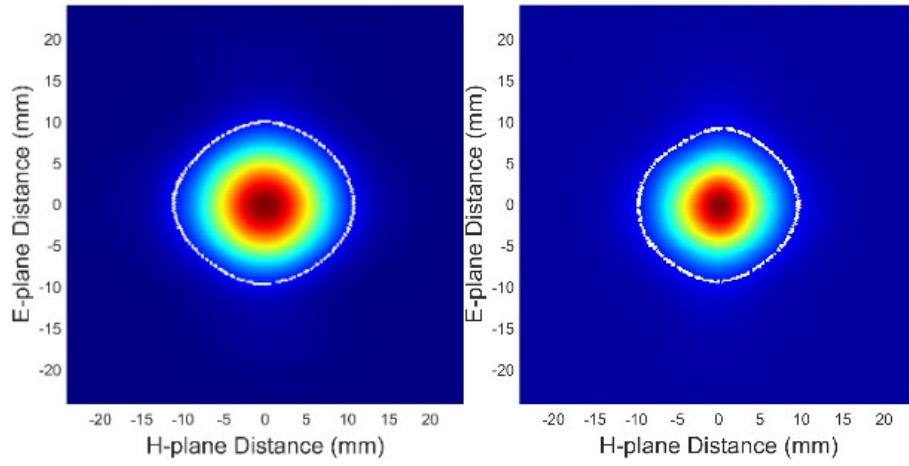
Camera image performance is investigated with two beams. First, at a fixed distance, the horn antenna directly illuminates the camera, with its diverging beam within the radiating near-field of the antenna. Second, the horn antenna feeds the 90° off-axis parabolic mirror to illuminate the camera with a near-collimated beam. With the latter, both cameras are set to acquire 2-D cross-sectional images of the beam at different distances along the propagation z -axis. Here, the single-pixel camera is chosen to acquire images with a resolution step size setting of $d_{px} = d_{py} = 200 \mu\text{m}$. Note that due to the nature of the raster scanning mechanism, the bespoke single-pixel camera requires anywhere between 3 to 6 hours of exposure time. In principle, having the same exposure time for the commercial camera would provide a fairer comparison. However, in this work, the commercial camera was configured to have an exposure time of around 1 minute to prevent overheating or damaging the system.

7.3.1 Diverging beam

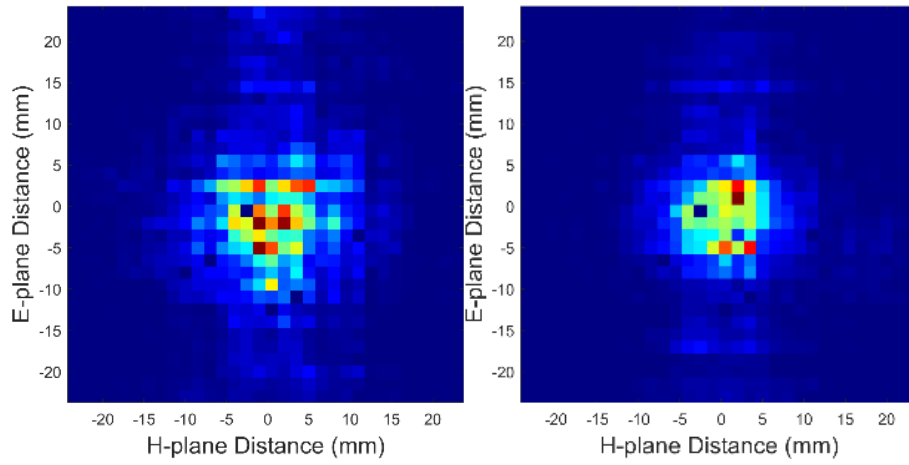
With the diverging beam, the distance between the antenna and camera apertures is fixed at 30 mm. Raw (unfiltered in this context) 2-D images showing the spatial beam profiles from electromagnetic (EM) simulation software (HFSS) and measured by the cameras, at 76 GHz and 92 GHz, are shown on Figure 7.3(a) to 3(c). In addition, images created using the commercial camera's proprietary median-smooth filter is applied (at the maximum level), shown in Figure 7.3(d). Also, MATLAB's 'disk smooth' filter [19] is applied to the raw images from the commercial camera and the results are shown in Figure 7.3(e).



(a)



(b)



(c)

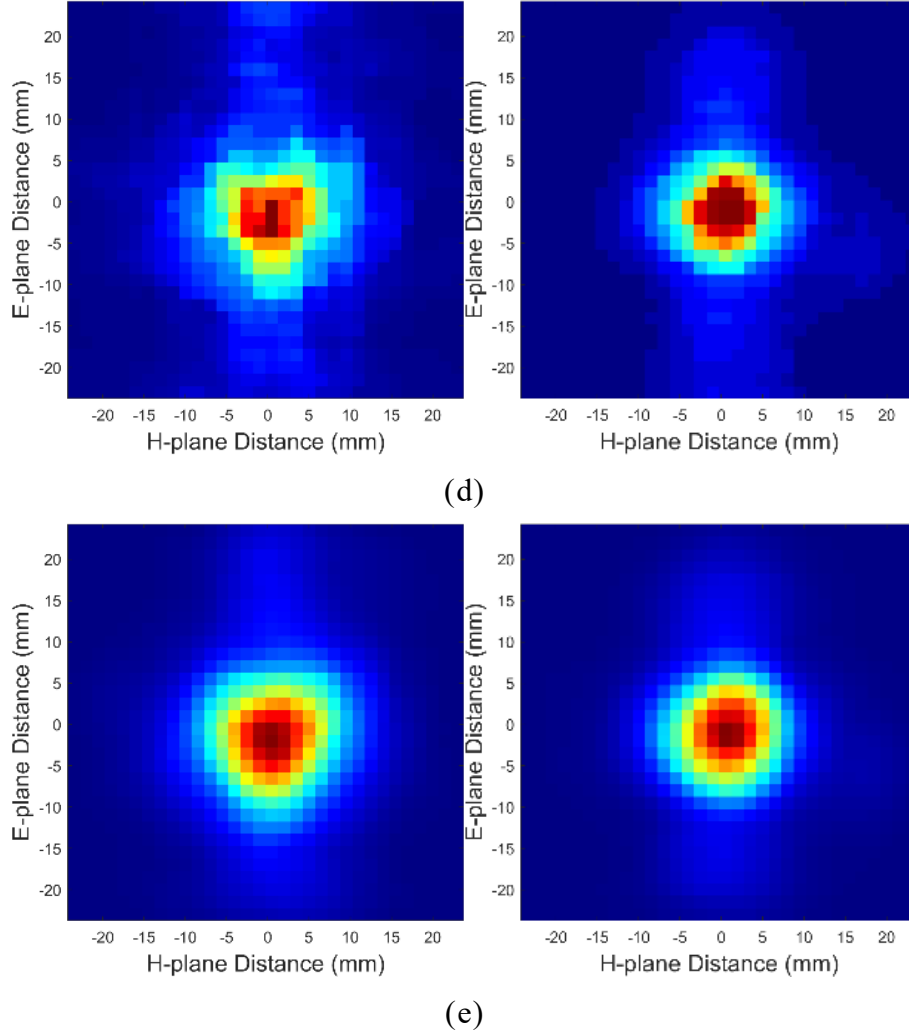


Figure 7.3 : EM Simulated and measured diverging beam cross section images (normalized values) at 76 GHz (left) and 92 GHz (right) with a 30 mm separation distance between antenna and camera apertures: (a) Reference HFSS (raw); (b) single-pixel camera (raw); (c) commercial (raw); (d) commercial (proprietary median-smooth filter); and (e) commercial (MATLAB ‘disk smooth’ filter). The white loci in 3(a)-3(b) represent the extracted beam boundaries.

From Figure 7.3, the rough shape and size of the incident beam can be estimated by visual inspection of the raw images from both cameras. However, with the commercial camera, the poor image resolution and fidelity prohibit extraction of useful information (e.g., cross-sectional beam profiles and beam widths). In contrast, with the single-pixel camera, the raw image can accurately identify the $1/e^2$ boundaries and beam centers. With smooth filters, the camera images resemble those simulated. Figure 7.3(d) and 3(e), clearly shows significant improvements on image quality for the beam shapes.

For example, beam cross-section profiles are plotted for the magnetic(H)- and electric(E)-field planes at 76 GHz and 92 GHz and the results are shown in Figure 7.4. Here, the full width at half maximum (FWHM) and $1/e^2$ beam widths are extracted and given in Table 7.2. It can be seen that the raw image from this single-pixel camera is in excellent agreement with simulations at both 76 GHz and 92 GHz. With the commercial camera, there is reasonable agreement with the use of its proprietary filter at both frequencies; the MATLAB filter gives a slightly better agreement with simulations.

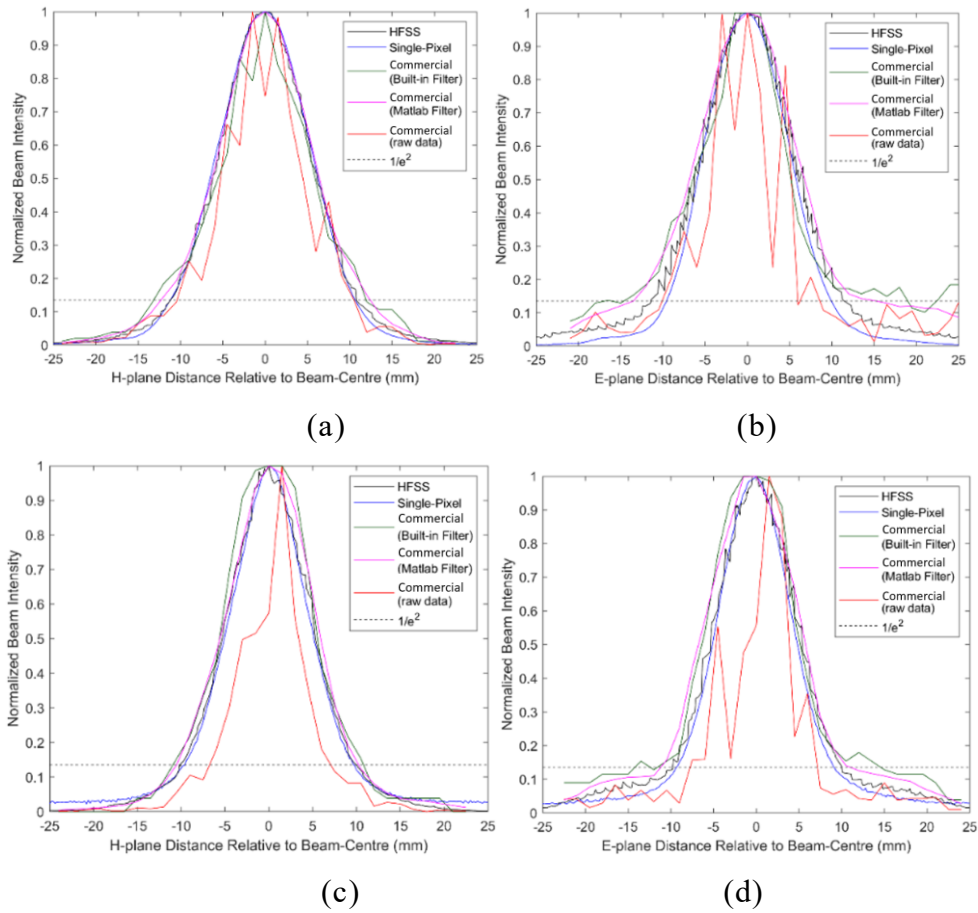


Figure 7.4 : Orthogonal beam cross-sectional profiles from reference HFSS simulations (raw), single-pixel (raw) and commercial camera images with 30 mm separation distance between the antenna and camera apertures: (a) H-field plane at 76 GHz; (b) E-field plane at 76 GHz; (c) H-field plane at 92 GHz; and (d) E-field plane at 92 GHz.

Table 7.2 : Beam widths extracted from measured images. The percentage difference in the extracted values from the camera image and EM simulated reference is given in brackets.

Camera Type	H-field Plane				E-field Plane			
	FWHM (mm)		1/e ² (mm)		FWHM (mm)		1/e ² (mm)	
	76 GHz	92 GHz	76 GHz	92 GHz	76 GHz	92 GHz	76 GHz	92 GHz
Reference HFSS Simulation (Raw)	12.4	10.6	23.0	20.1	12.3	10.7	22.9	19.8
Single-pixel Camera (Raw)	12.4 (0.4)	10.2 (3.8)	21.8 (4.9)	19.3 (4.0)	11.7 (4.9)	10.0 (6.5)	19.5 (14.8)	18.5 (6.6)
commercial (Proprietary Filter)	11.5 (7.0)	11.2 (5.7)	25.1 (9.3)	21.8 (8.5)	11.6 (5.8)	11.5 (7.4)	33.6 (46.8)	26.2 (32.3)
commercial (MATLAB Filter)	12.7 (2.9)	11.5 (8.5)	26.7 (16.4)	20.8 (3.5)	13.7 (10.7)	12.5 (17.2)	28.7 (25.4)	22.2 (12.1)

Caution should be taken when extracting beam profile information using the commercial camera, even when using a smooth filter, due to poor image quality. Moreover, while this commercial camera is specified for operation down to 50 GHz, the difference performance for the 1/e² beamwidths in Table 7.2 is noticeably worse at 76 GHz, when compared to with the narrower beam 92 GHz.

A reference analytical model is used to evaluate how accurately each camera captures H- and E-field plane beam cross-section intensity distributions, in terms of overall shape, using statistical *R*-squared analysis. The analytical model assumes that the pyramidal horn antenna acts as a perfect Gaussian source, with a Gaussian approximation for beam intensity given by [20]:

$$I(r) = I_0 \cdot e^{-2\left(\frac{r}{w_z}\right)^2} \quad (7.1)$$

where I_0 is peak beam intensity (or irradiance); r is the radial distance from the center axis of the beam; and w_z defines the beam radius at $1/e^2 \sim 13.5\%$ of its peak intensity (or $1/e$ of its peak field). From Table 7.2, the beam width $2w_z$ is taken from EM simulations and $I(r)/I_0$ is calculated against r , such that this analytical model can be compared directly against the plots in Figure 7.4, to extract the R -squared values. The results are given in the bar chart shown in Figure 7.5. Both the EM simulation and single-pixel camera images show a good fit to the Gaussian approximation for beam intensity. As expected, with the commercial camera, there is reasonable agreement with the use of its proprietary filter at both frequencies; the MATLAB filter gives a slightly better agreement with simulations.

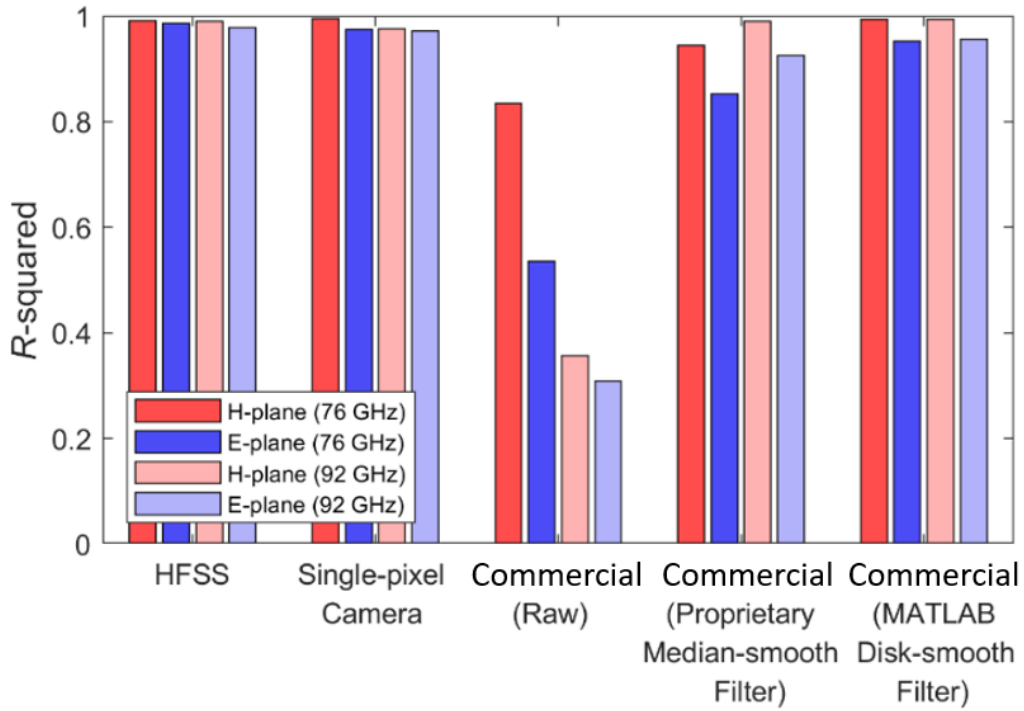


Figure 7.5 : R -squared values for beam intensity profiles against the Gaussian approximation reference analytical model.

7.3.2 Collimated Beam

With the diverging-beam, the peak beam intensity is relatively high (due to the short separation distance) and the $1/e^2$ beamwidth is only approximately 44% of the commercial camera aperture size, resulting in a large signal-to-noise ratio (SNR).

In this subsection, a collimated beam is used having a more complex profile and in a more challenging SNR environment. With an ideal single-point source and mirror (optical) assembly, geometric ray tracing shows that only one axis of symmetry exists (unlike the orthogonal two-axis symmetry for the diverging beam). Moreover, with the horn-mirror (quasi-optical) assembly, there will be additional distortion to the beam cross-section to that found with the optical assembly. In addition, even small mechanical misalignments will add further distortion to the beam's cross-sectional profile.

With the same output power levels, the total separation distance is now increased from 30 mm to 246/346 mm and the collimated $1/e^2$ beamwidths are now of the same order as the commercial camera aperture size; resulting in a significantly degraded SNR. Moreover, with these larger beam widths, knife-edge diffraction from the front-facing aluminum casing around the aperture may cause further beam image distortion. Also, with any collimated beam, the effects of standing waves will become more pronounced from a camera having a high reflectance aperture. Two sets of images are taken at 76 GHz and 92 GHz, at different distances (170 mm and 270 mm) from the mirror along the propagation z -axis.

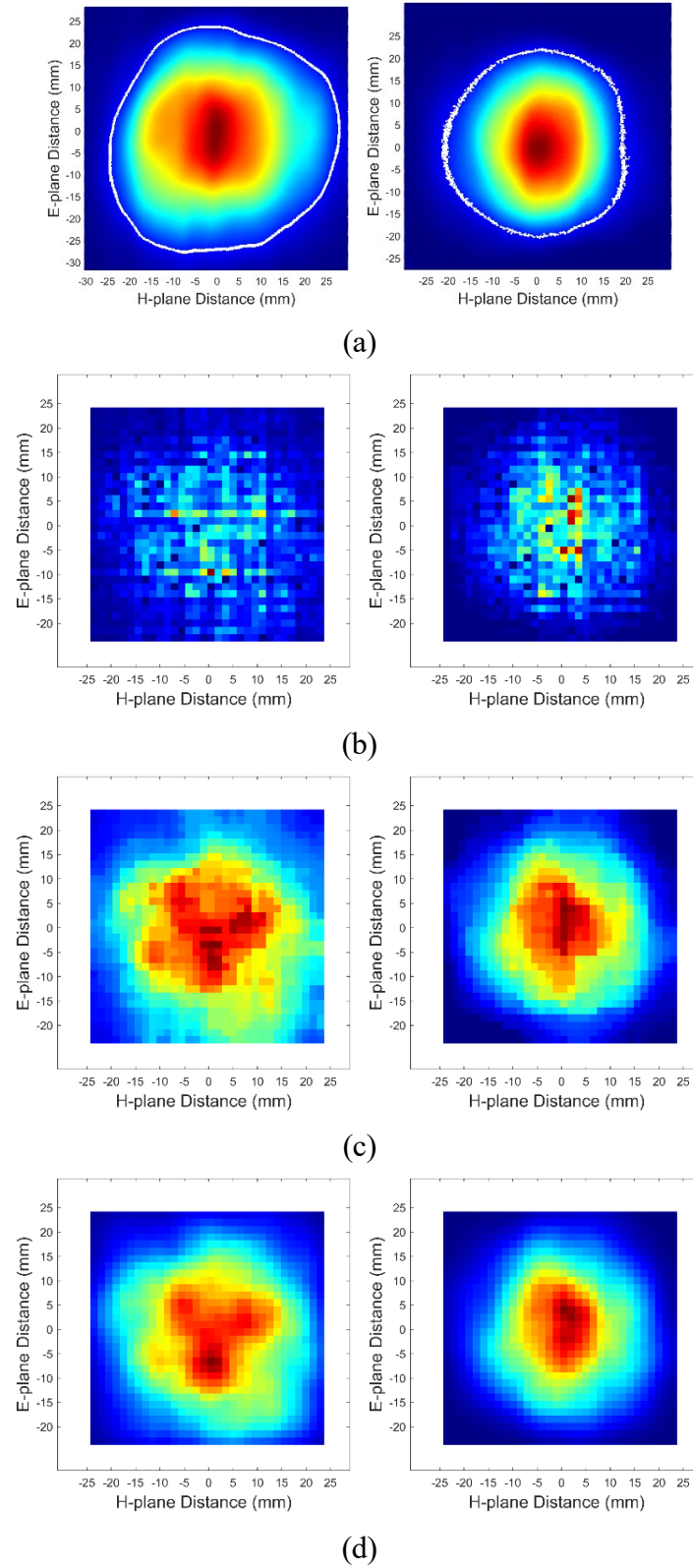


Figure 7.6 : Measured collimating beam images (normalized) with a 170 mm separation distance between mirror center and camera aperture at 76 GHz (left) and 92 GHz (right): (a) single-pixel (raw); (b) commercial (raw); (c) commercial (proprietary median-smooth filter); and (d) commercial (MATLAB ‘disk smooth’ filter). The white loci represent the extracted beam boundaries.

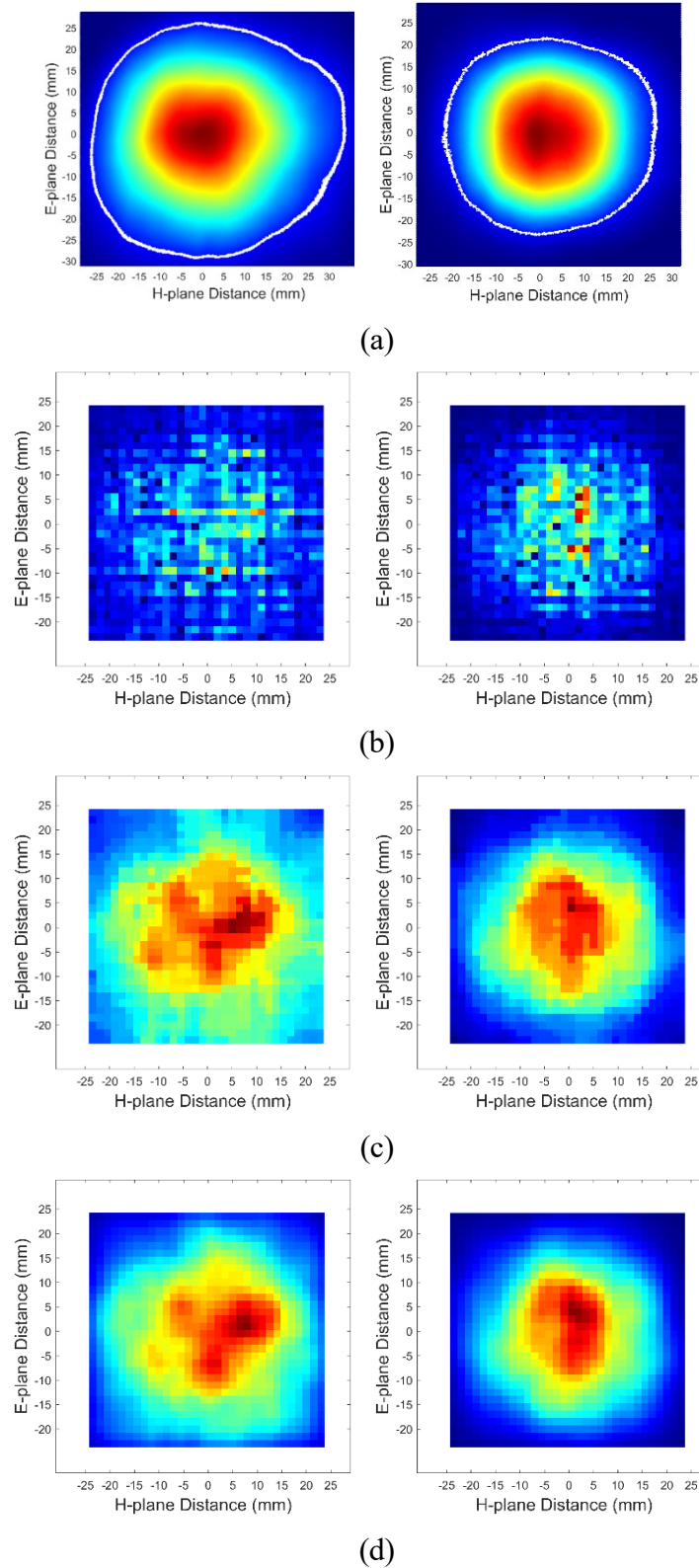


Figure 7.7 : Measured collimating beam images (normalized values) with a 270 mm separation distance between mirror center and camera aperture at 76 GHz (left) and 92 GHz (right): (a) single-pixel (raw); (b) commercial (raw); (c) commercial (proprietary median-smooth filter); and (d) commercial (MATLAB ‘disk smooth’ filter). The white loci represent the extracted beam boundaries.

Figure 7.6 and Figure 7.7 show the resulting images taken at separation distances of 170 mm and 270 mm, respectively. The single-pixel camera clearly shows the expected distortion in the beam cross-section; as with the diverging beam, the $1/e^2$ beam widths at 92 GHz are noticeably smaller than at 76 GHz. In contrast, in this low SNR environment, the random noise adds to the existing problems of poor resolution and pixel quality, resulting in the commercial camera being unable to resolve the complex image seen by the single-pixel camera. For example, using only the raw image from the single-pixel camera, it can be seen that the $1/e^2$ beam widths slightly increase ($\sim 10\%$ in the H-field plane and $\sim 1\%$ in the E-field plane) as the separation distance increases from 170 mm to 270 mm; demonstrating that the beam is not perfectly collimated and that this assessment cannot be made with the commercial camera.

By observation of Figure 7.6(c), 6(d), 7(c) and 7(d), pixels with high-intensity values seen in raw commercial camera images are randomly distributed across the array. As a result, after applying the smooth filter, it becomes impossible to locate the beam center.

7.4 Standing-wave Detection

With the commercial camera, significant periodic fluctuations in the detected pixel power levels are observed when the camera is moved along the propagation z-axis. This phenomenon indicates the presence of standing waves between the camera and the source (also observed with the single-pixel camera, but to a lesser degree).

With a collimated beam, the spatial power intensity distribution is measured with both cameras. Each camera is moved along the propagation z-axis; 150 mm to 270 mm from the center of the 90° off-axis parabolic mirror. With the single-pixel camera, RAM is always used to minimize standing waves. Therefore, to observe its effectiveness, measurements are undertaken with and without the RAM. With the commercial camera, only the central pixel is chosen to record the power intensity.

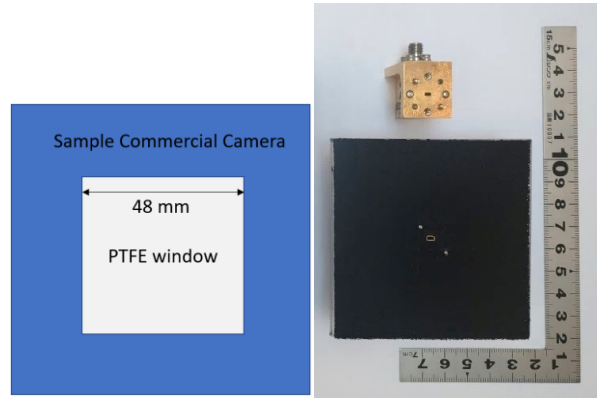


Figure 7.8 : Illustration of the commercial camera with $48 \text{ mm} \times 48 \text{ mm}$ aperture (left) and single-pixel camera detectors with $2.54 \text{ mm} \times 1.27 \text{ mm}$ aperture (right) without (top) and with (bottom) RAM.

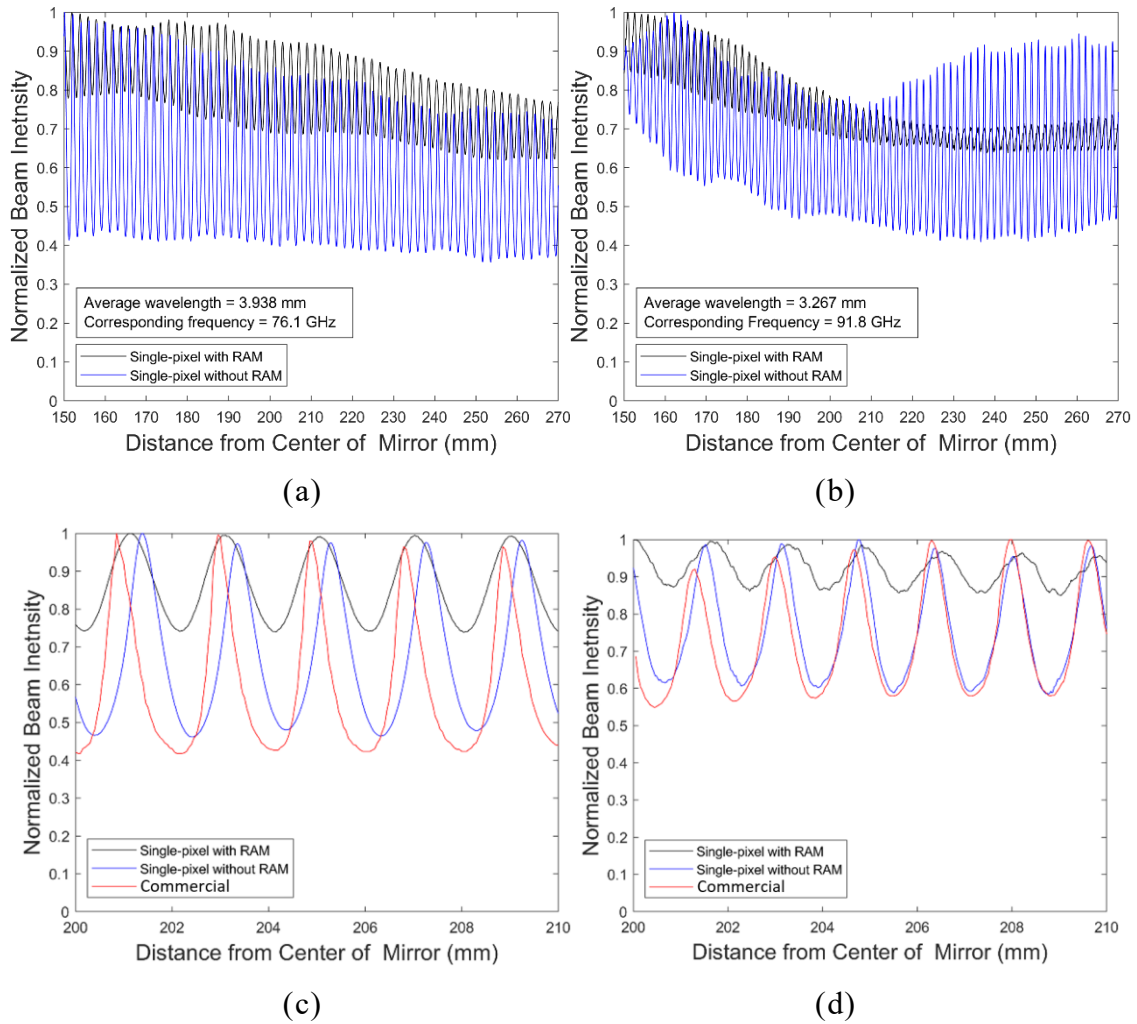


Figure 7.9 : Beam intensity along the propagation z -axis (running average with 11 points): (a) Single-pixel at 76 GHz; (b) Single-pixel at 92 GHz; (c) Both camera at 76 GHz (zoomed-in); and (d) Both cameras at 92 GHz (zoomed-in).

Figure 7.8 shows the apertures for both cameras and Figure 7.9(a) and 9(b) give the normalized propagation z -axis intensity distributions for the single-pixel camera with and without the RAM at 76 GHz and 92 GHz, respectively. It can be clearly seen that intensity has sinusoidal-type ripples along the propagation z -axis. At 76 GHz, the average distance between two peaks is measured as $\lambda_o/2 = 1.969$ mm, corresponding to a frequency of 76.1 GHz. Similarly, at 92 GHz, the average distance between two peaks is measured as $\lambda_o/2 = 1.634$ mm, corresponding to a frequency of 91.8 GHz. The RAM surrounding the aperture of the open-ended waveguide reduces the power intensity of the standing wave by more than 3 dB. Figure 7.9(c) and 9(d) show normalized propagation z -axis intensity distributions for both cameras, over a zoomed-in range of 10 mm. At both 76 GHz and 92 GHz, the commercial camera show more prominent standing waves, when compared to the single-pixel camera with RAM and is of the same order without RAM.

The voltage standing-wave ratio (VSWR) can, in principle, be read-off directly from Figure 7.9. However, any extracted value for VSWR is meaningless since the source is also not impedance matched to that of the reference (impedance of free space η_o in this QS case). This will be found even with the most expensive commercially available vector network analyzers (VNAs), typically used as the source and detector, due to non-ideal components employed within the test set (irrespective of whether error correction is applied).

7.5 Conclusion

Mm-wave camera allows end-users to perform various types of characterization very conveniently. Therefore, it is one of the most useful measurement tools for characterizing 3-D printed quasi-optical components. To trust the results that a commercial mm-wave camera produces, it is important to benchmark its performance and identify the advantages and disadvantages of using the camera. Here, a sample commercial camera was benchmarked against a custom-built single-pixel camera at W-band, for image quality and aperture reflectance.

When compared to a reference single-pixel camera, the commercial camera exhibits limited image resolution and fidelity, with significant levels of systematic spatial noise (due to faulty pixels and high pixel-to-pixel responsivity deviation). In a poor signal-to-noise scenario, the addition of random noise exacerbates the problems with poor resolution and pixel quality. While the commercial camera technology can provide a convenient means of measuring beam profiles at room temperature, in low SNR environments poor image quality prohibits accurate characterization. It was identified to be possible causes of both beam and image distortion; this gives important insight into the best use of (sub-)THz cameras and interpretation of their images.

While the custom-made non-pre-amplified direct-detection single-pixel camera provides excellent image resolution and fidelity, it inherently suffers from very slow raster-scanning speeds and operational bandwidth limitations. For this reason, the commercial camera technology is excellent for performing qualitative measurements in real-time, with the caveats outlined in this work. In addition, this work has shown that the bespoke single-pixel camera is an excellent tool for quasi-optical imaging, which can be used for characterizing and inspecting 3-D printed mm-wave components such as lenses, mirrors, and zone plate.

7.6 Acknowledgements

Work presented in this chapter was published on the paper: “Benchmarking a commercial (sub-)THz focal plane array against a custom-built millimeter-wave single-pixel camera” by **S.-H. Shin**, and S. Lucyszyn, (*IEEE Access*, vol. 8, pp. 191174 - 191190, Oct. 2020) [1]. The research on this work has been performed under the supervision of Prof. S. Lucyszyn and he went over every word of the original article.

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8 Conclusion

The research objective of this PhD thesis is to investigate the application of various consumer-level polymer-based 3-D printing techniques for fabricating microwave and mm-wave components and subsystems. Polymer-based 3-D printing offers much faster, cheaper, more flexible prototyping and bespoke small batch production compared to conventional manufacturing techniques. As many microwave and mm-wave components are highly specialized and produced on demand, 3-D printing technologies are the ideal candidate. By comparing the advantages and disadvantages of each technique, it was possible to identify an appropriate 3-D printer for each component. Both properties of the 3-D printers and building material were considered to produce an optimal result. Bespoke printing and post-processing procedures appropriate for each component were meticulously considered and introduced. Following these processes, various 3-D printed proof of concept components such as Ku-band phased-array, G-band dielectric lenses, polarizers, and mirrors were produced. In addition, to accommodate the highly customized nature of each component, tailored testing platforms were 3-D printed for characterizing processes. A multi-channel quasi-optical subsystem was also printed to demonstrate the integration of several different types of components. Finally, a custom-made single-pixel mm-wave camera was developed for future use on characterizing and defect-finding 3-D printed quasi-optical components.

The first chapter summarized general trends in current polymer-based 3-D printing for microwave and mm-wave applications. A literature survey of components relevant to this dissertation is included. In Chapter 2, general information, including characteristics, printing mechanism, advantages and disadvantages of the most common polymer-based 3-D printing techniques, were discussed. Advantages and disadvantages of printing techniques such as FDM, SLA, MSLA and material jetting were presented and compared.

In Chapter 3, detailed set-ups for printing and post-processing techniques that were adopted and developed for this work were presented. Selection criteria of critical set-up parameters such as layer thickness and build orientation were discussed in this chapter. In addition, possible safety hazards of each 3-D printer were identified. Finally, the

process of setting-up a safe 3-D printing fabrication and post-processing laboratory environment were presented. An example of the lab set-up was given with detailed information on selecting, purchasing and/or building appropriate safety equipment.

In Chapter 4, a combination of commercial-level material jetting and a low-cost desktop FDM printer was used to fabricate a fully 3-D printed 4-elements phased-array operating at Ku-band (12 to 18 GHz). By integrating 26 waveguide-based components ($1 \times$ flange, $6 \times$ mitred bends, $3 \times$ power splitters, $4 \times$ tunable delay lines, $4 \times 90^\circ$ waveguide twists, $4 \times$ bent interconnects and $4 \times$ H-plane sectoral horn antennas) as a two-piece split-block structure, potential of using 3-D printing a geometrically complex multi-component structure was shown. The phased-array antenna showed outstanding measured performance having tunable steering angles of 54° and 52° at 15 GHz and 17 GHz, respectively.

Moving up in frequency, in Chapter 5, the design, fabrication and measurement process of the quasi-optical component were covered; horn antennas, mirrors, RAM, grid polarizer, and dielectric lenses were presented for G-band (140 to 220 GHz) applications. In addition, A simple 3-D printed quasi-optical subsystem was employed to be used for integrating and testing the quasi-optical components. Using the quasi-optical subsystem, it was possible to characterize each component accurately and rapidly without having to spend a significant amount of time on measurement set-up and components alignment. Measured results showed that most of the quasi-optical components have comparable performances as their commercial counterparts despite a fraction of fabrication cost.

Chapter 6 follows up on the work from Chapter 5 and demonstrates a multi-channel subsystem that integrates various quasi-optical components. Some components were adopted or redesigned based on the work in Chapter 6, while components such as convex-lens, holey beamsplitter and bespoke quasi-optical test platforms were newly fabricated to be adopted for the final subsystem. The final subsystem showed the potential of a fully 3-D printing functional front-end subsystem consisting of a relatively complex quasi-optical beam network and corresponding components.

Finally, in Chapter 7, an mm-wave raster scanning single-pixel camera was custom built. It was used to benchmark a commercial THz camera which was one of the main tools used for characterizing various 3-D printed quasi-optical components such as lens

and mirror. The custom-built mm-wave is planned to be used for future studies on characterizing or finding 3-D printing defects on various quasi-optical components such as zone-plates, graded index lenses and mirrors.

Despite the limitations caused by the current 3-D printing technologies that were employed in this thesis, this research work has proved that additive manufacturing technologies have great potential to be used for rapid prototyping and on-demand production of low-cost microwave and mm-wave components. Moreover, it has been shown that 3-D printing an entire subsystem significantly reduces fabrication time, cost and measurement complexities. With the continuous advancement in additive manufacturing technologies and building material, as well as more studies on characterization methods of the printed components, 3-D printing is expected to be adopted more and more by academia and industries as a standard microwave and mm-wave development tool.

Publications

Journal Articles

1. **S.-H. Shin**, D. Alyasiri, M. D'Auria, W. J. Otter, C. W. Myant, D. Stokes, Z. Tian, N. M. Ridler, and S. Lucyszyn, "Polymer-based 3-D printed Ku-band steerable phased-array antenna subsystem," *IEEE Access*, vol. 7, pp. 106662-106673, Aug. 2019.
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4. **S.-H. Shin**, X. Shang, N. M. Ridler, and S. Lucyszyn, "Polymer-based 3-D printed 140-220 GHz low-cost quasi-optical components and integrated subsystem assembly", *IEEE Access*, vol. 9, pp. 28020-28038, Feb. 2021.
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6. **S.-H. Shin**, R. Payapulli, L. Zhu, M. Stanley, X. Shang, N. M. Ridler, and S. Lucyszyn, "3-D printed plug and play prototyping for low-cost sub-THz subsystem", *IEEE Access*, Accepted for Publication

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1. **S.-H. Shin**, D. Alyasiri, M. D'Auria, W. J. Otter, C. Myant, D. Stokes, Z. Tian and N. M. Ridler, S. Lucyszyn, "Fully 3-D printed tunable microwave subsystem," *IEEE MTT-S International Microwave Workshop Series on Advanced Materials and Processes (IMWS-AMP)*, Bochum, Germany, Jul. 2019 (**Best Student Paper Award**).