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APPLICATIONS

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This dissertation is dedicated to my wonderful parents, who raised the first-generation Ph.D. scholar in the entire family, sharing their visionary support and guidance throughout the entire journey. May Allah (SWT) be pleased with you and have mercy on you in both worlds (Ameen)

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Abstract

The success of future multi-functional communication, radar, sensing, and countermeasure systems heavily relies on ultrawideband (UWB) antennas. UWB antennas have a very high demand in electromagnetic compatibility (EMC) testing, radar calibration, satellite tracking, reflector feeds, remote sensing, wideband instrument for snow measurements (WISM), medical imaging, through-wall imaging, and UWB radar systems. In an ideal case, these multi-faceted systems require the use of a single UWB “one-size-fits-all” antenna that satisfies the bandwidth and radiation requirements compared to using multiple narrowband antennas saving time, labor, and overall cost of the system.

This dissertation delves into the theory, design, and feasibility of one of the kinds of a single UWB antenna or a single integrated UWB antenna system for their potential use in the characterization or calibration of weather radars or other UWB communication systems. Firstly, detailed Maxwell-Garnett particle mixing analytical models are presented and evaluated based on their performance and error analysis. The correct analytical model is identified and used in the synthesis of tunable artificial dielectric material (TADM) at microwave (S-band) and mm-wave (W-band) frequencies for applications such as RF substrates for antennas and microwave components such as filters. This technique allows for the utilization of cost-effective materials that already offer outstanding thermal properties and seamless compatibility with a multi-layer fabrication process, can now be equipped

with desired tunable ε_{eff} and loss tangent characteristics. The measured results show substantial reductions of 45% in ε_{eff} and 38% in loss tangent values in the S-band, along with reductions of 32% and 72% in ε_{eff} and $\tan\delta$, respectively, in the mm-Wave frequency band. Impressively, the simulated, calculated, and measured results demonstrate an excellent level of agreement.

To see the feasibility of using commercially available state-of-the-art UWB antennas for in-situ characterization of radars, a detailed analysis is carried out based on the idea of using a single UWB probe antenna for radars operating in the L-, S-, C-, X-, Ku-, and K-bands. This analysis shows in detail that using commercially available UWB antennas is not suitable for in-situ measurements due to their wide beamwidth performance, especially for weather radars operating in the most commonly used S-band at 3 GHz. To solve this challenge, several solutions are presented. Among the solutions presented based on using corrugated horns and multi-layer or hyperbolic dielectric lenses, the Maxwell-Garnett theory that laid the foundation of the design of tunable artificial dielectrics is also applied in the synthesis of a UWB graded-index (GRIN) artificial dielectric lens antenna (ADLA). Proposed novel solutions make the use of these commercially available antennas viable for radars and communication systems.

A UWB open-boundary dual-polarized quad-ridged horn antenna (QRHA) having an impressive absolute bandwidth of more than 30 GHz (32:1) covering 1-32 GHz, is then presented making it an ideal choice for UWB communication and radar systems covering a wide range of applications relying on common frequency bands including L, S, C, X, Ku, and K. The unique design allows for a favorable impedance matching, acceptable stable and narrow-beamwidth performance, and most importantly no-pattern degradation at higher frequencies which is a common concern for most ridged-horn antennas designed thus far.

Chapter 1

Introduction

The rapid advancement of radars, communication systems, and multi-functional RF front-end systems has generated a serious need for ultrawideband antenna systems. Several antennas are usually used across different frequency bands to achieve multiple functionalities. However, with the use of an ultrawideband (UWB) antenna system, multiple applications can be targeted using one single antenna aperture saving cost, weight, size, time, labor, and power consumption. The greatest benefit of a UWB communication system is evident from the famous Shannon equation which relates channel capacity ($\mathbf{C}$) to the bandwidth ($\mathbf{B}$) as given by (1.1) [1]:

$$\mathbf{C} = \mathbf{B} \log \left(1 + \frac{P}{\mathbf{B} \times N_0} \right) \quad (1.1)$$

Where, $\mathbf{C}$ is the channel capacity in bits per second (bps), $\mathbf{B}$ is the channel bandwidth in Hz, P is the power of the signal in Watts, and N_0 is the noise power spectral density in Watts/Hz. This equation shows that for the given power P , capacity ($\mathbf{C}$) can increase with bandwidth ($\mathbf{B}$).

UWB antennas have been widely used in communication system applications

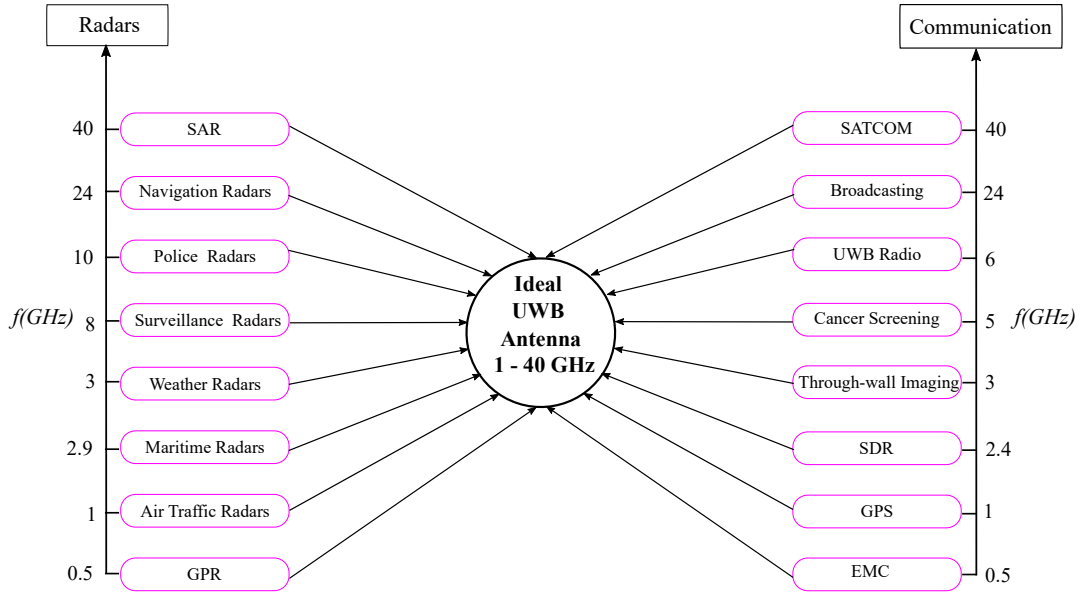


Figure 1.1: Conceptual diagram of the applications of an ideal UWB antenna in both radars and communication systems.

such as EMC testing, satellite tracking, reflector feeds, and other detection systems [2–5]. They have also been used in applications related to remote sensing and synthetic aperture radar (SAR) based snow measurements [6], through-wall imaging [7], bio-medical imaging [8], cancer screening [9], automobile collision avoidance [10], software-defined radio [11], UWB radio systems [12, 13], and UWB radars [8]. Figure 1.1 shows a conceptual diagram of the possible applications of a UWB antenna system in both radar and communication systems having an ideal bandwidth of 1-40 GHz.

1.1 Problem Statement

1. **Non-existence of a UWB antenna probe for Unmanned Aerial Vehicle-based in-situ radar characterization:** The studies reported in [14–17] related to probe selection for a Unmanned Aerial Vehicle (UAV)-based radar characterization have selected a “Stop Sign: SS” microstrip patch antenna array of 4×4 elements whose geometry is shown in Figure 1.2, and a horn antenna element. *However, one key limitation to these probes is the narrow bandwidth performance. For instance, in the case of the microstrip SS array, the bandwidth is merely 70 MHz, while in the case of the horn, it is around 100 MHz, both targeting the S-band. This narrow bandwidth performance restricts the use of such probes across multiple frequency bands. Moreover, each antenna behaves differently in terms of its RF performance, when mounted on the UAV. This further involves the utilization and subsequent re-calibration of multiple probes, thereby contributing to elevated costs and increased labor efforts. Thus a single UWB antenna is highly desired that when calibrated with the drone, can be used for radar characterization in covering several frequency bands.*
2. **Non-existence of an acceptable Quad-ridged Horn Antenna:** As will be discussed in the coming chapter that the Quad-ridged Horn Antenna (QRHA) is the ideal candidate for not only UWB communication systems but also for UWB radar systems. However, these antennas suffer from pattern degradation at frequencies usually greater than 12 GHz due to the presence of higher-order modes. This significant degradation makes these antennas unsuitable for UWB communication or radar systems where “well-matched patterns” are required. These antennas have very wide beamwidth performance typically 150° – 80° es-

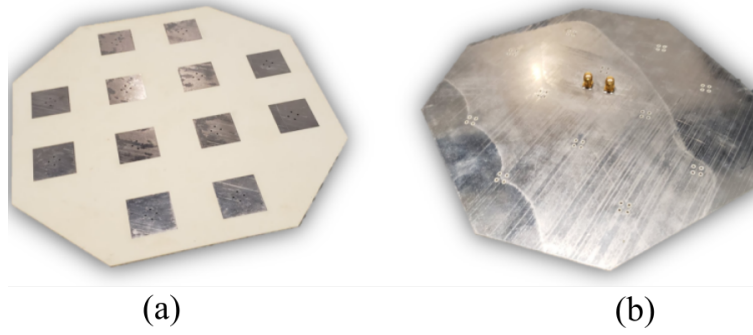


Figure 1.2: The microstrip SS patch array having a bandwidth of only 70 MHz (a) front view (b) rear view [17].

pecially in the lower part of the frequency band from 1-6 GHz, where 150° corresponds to 1 GHz and 80° corresponds to 6 GHz. The wide beamwidth performance exhibited by these antennas restricts their use in a crucial segment of the frequency band, which is predominantly utilized by weather radars and communication systems. It will be shown through detailed analyses in Chapter 2 that even the commercially available probe antenna in [18] is not suitable for UAV-based radar calibration due to its wide beamwidth performance in the lower region of the band where most of the weather radars operate. Such wide beamwidth radiation performance creates undesired scattering from the UAV structure thus producing ripples in the co-polar patterns as well as affecting the cross-polarization performance. Given these challenges, the majority of research on ridged-horn antennas done in the past is centered around frequencies starting from 6 or 8 GHz and beyond. *Thus, there is the non-existence of a suitable UWB antenna that covers for instance the 1-32 GHz band without pattern degradation and at the same time having good VSWR along with narrow and stable beamwidth performance throughout the entire band of operation.*

1.2 Motivation

The preceding discussions have culminated in the identification of a research problem that needs to be addressed. The motivation behind this endeavor is to have the following goals:

1. To assess, analyze, and compare several Maxwell-Garnett particle mixing analytical models and to choose the correct model through detailed error analyses. It is then desired to use this technique in the synthesis of tunable artificial dielectric material (TADM) at microwave (S-band) and mm-wave (W-band) frequencies for applications such as RF substrates for antennas and microwave filters. The analytical model must also be validated and compared for its performance with full-wave simulation, and measured results. (This work constitutes Chapter 4).
2. To use the particle mixing theory in the design of a 3D-printed graded index (GRIN) dielectric lens to address the wide beamwidth problem in the existing UWB QRHAs, aimed at in-situ characterization of weather radars using UAVs. Aside from this, it is also desired to propose other efficient and cost-effective solutions to the aforementioned problem. Specifically, the desired UWB antenna must satisfy the probe requirements set by previous works for the UAV-based characterization of radars, i.e. along with UWB performance, the desired probe must also have narrow and stable 3 dB HPBW performance of 40° in both E- and H-planes, and co-polar mismatch of no greater than 0.1 dB in H- and V-polarized patterns, in the entire frequency band of operation. (This work forms Chapter 5).
3. To present a low-cost and simple UWB antenna to meet the requirements of

both UWB communication and radar systems. A single UWB “one-size-fits-all” antenna covering the 1-32 GHz band, is thus highly desired covering the important radar frequency bands such as L-, S-, C-, X-, Ku, and K-bands. Such a UWB antenna can eradicate the necessity for re-installation and re-calibration of multiple antennas that are usually required to cover various frequency bands, thus providing substantial savings in terms of cost, time, and labor. Along with the UWB and narrow beamwidth performance in the 1-6 GHz region, it is highly desired that the proposed antenna shows good voltage standing-wave ratio (VSWR) performance ($VSWR < 2$), especially in the 1-6 GHz region, and at the same time is free of pattern degradation above 12 GHz thus addressing major challenges that are currently associated with the commercially available as well as existing QRHA literature works. (This work forms Chapter 6).

1.3 Hypothesis

This research hypothesizes that lenses based on artificial dielectric material can address the current challenges related to wide beamwidth and beamwidth variability across the entire band when using commercially available ridged-horn antennas for UAV-based radar characterization. It is also hypothesized that the dual-polarized QRHA is a “one-size-fits-all” candidate that is capable of providing a low-cost and efficient UWB performance spanning the 1-32 GHz band, thus covering important radar and communication system frequency bands. Lastly, this research postulates that the proposed QRHA is capable of addressing major challenges currently faced in the radar community in terms of wide beamwidth performance, pattern degradation, and poor impedance matching performance.

1.4 List of Contributions

The main contributions originated from this research work are the following:

1. For the first time, this research work presents a detailed study, compilation, comparison, and assessment of several analytical models related to the Maxwell-Garnett particle mixing theory. Through full-wave simulations and detailed comparison as well as error analyses, the correct analytical model is identified. For the first time, the effects of the direction of excitation as well as the polarization of the incident field on the ε_{eff} of the anisotropic particle are analyzed and discussed in detail.
2. Particle mixing theory and the correct analytical model are used in the synthesis of a low-loss tunable artificial dielectric material (TADM) for antennas and microwave components. The analysis is validated by fabricating and testing the proposed TADM in the microwave frequency (S-band) at 3.5 GHz, as well as in the mm-Wave frequency (W-band) at 77 GHz. These contributions together resulted in [19].

The aforementioned contributions establish the groundwork for a solution that can tackle the losses associated with the silicon-based semiconductor fabrication of on-chip antennas. By using TADL, ε_r and $\tan\delta$ of the portion of the silicon substrate on which the antenna is to be realized, can be lowered to improve the efficiency and gain of the on-chip antenna. Moreover, this technique has the capability of mitigating the scan blindness resulting from the surface waves [20].

3. Through rigorous analyses and far-field measurements of the commercial QRHA QH2000 probe antenna [18] including the UAV platform, the incompat-

ibility of such QRHAs is reported due to the wide beamwidth performance of the QRHA. This contribution led to several important findings. Based on this study it is concluded that commercially available QRHA generates undesired scattering from the drone structure, leading to the development of undesired ripples in the co-polar radiation patterns and high levels of cross-polarization. This study contributed to [21–23].

4. To address the wide beamwidth problem of the commercially available QRHAs for UAV-based in-situ characterization of radars, for the first time, several solutions are presented including the design of a 3D-printed graded index (GRIN) dielectric lens [24], multi-layer superstrate lens [25], hyperbolic lens [26], and corrugated QRHA [21].
5. For the first time, a “one-size-fits-all” low-cost and simple UWB antenna covering a 1-32 GHz band, is proposed to satisfy the requirements of both UWB communication and radar systems. The proposed antenna covers the important frequency bands such as L, S, C, X, Ku, and K. Along with the UWB and narrow beamwidth performance in the lower band: 1-6 GHz, the proposed QRHA exhibits good VSWR performance ($VSWR < 2$), especially in the 1-6 GHz region, and at the same time is free of pattern degradation above 12 GHz thus addressing major limitations of QRHAs. This work contributes to [27, 28].

1.5 Dissertation Overview

This dissertation is divided into seven chapters. This chapter (Chapter 1) highlighted the problems associated with the current UAV-based radar characteri-

zation setup as well as discussed the motivation and main contributions of this work.

Chapter 2 provides a detailed literature review as well as a comparison of UWB antennas including both arrays and standalone antennas. It also highlights the choice of the antenna probe for UAV-based radar characterization. Furthermore, this chapter provides an overview of the past in-situ calibration techniques and highlights the RF requirements for UAV-based radar characterization.

Chapter 3 identifies and addresses the incompatibility of the commercially available ridged-horn antenna probes for UAV-based in-situ characterization of radars due to their wide beamwidth performance, especially in the lower frequency band.

Chapter 4 provides an in-depth review as well as a comparison of several particle mixing analytical models. It validates the proposed particle mixing technique through detailed error analyses, simulations, and measurements of several artificial dielectric samples at microwave and mm-wave frequencies.

Chapter 5 presents and discusses several dielectric lens antenna solutions as well as a corrugated ridged horn antenna to address the wide beamwidth problem of commercial ridged horn antennas. A “one-size-fits-all” QRHA UWB antenna covering the 1-32 GHz band is presented in Chapter 6 that satisfies the requirements of both UWB communication and radar systems. Finally, Chapter 7 summarizes the findings of this dissertation, presents its conclusions, discusses limitations, and outlines directions for future research.

Chapter 2

Literature Review and Background

This chapter provides an overview of UWB antenna arrays as well as standalone antenna elements. Limitations in terms of cost and complexity vs absolute bandwidth are discussed for both UWB arrays and single antenna elements. The choice of the antenna element between arrays and standalone antenna is contingent upon the targeted application. For instance, one can choose a UWB array consisting of several elements if electronic beam-steering is required. It should be noted that for UAV-based in-situ measurements, beam steering is not required. Thus, standalone antennas can be chosen for this task if: firstly, they can provide UWB performance so that a single probe UAV platform can be used for the characterization of several radars targeting S-, C-, X-, Ku-, and K-bands, secondly, they satisfy the UAV-based probe beamwidth requirements. It will be discussed and shown in the following chapters that a standalone antenna such as QRHA is preferred over arrays for UAV-based radar characterization. This choice is driven by the standalone antenna's ability to offer a broader absolute bandwidth compared to arrays while maintaining a similar narrow beamwidth performance with a relatively simpler geometric configuration.



Figure 2.1: Tightly-Coupled Dipole Array having frequency range between 2-18 GHz (a) single-polarized (b) dual-polarized [32].

2.1 UWB Antenna Arrays

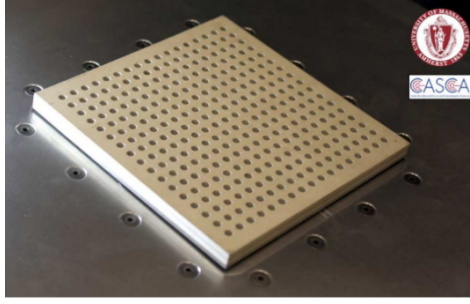
Based on their configuration, UWB arrays can be categorized into three groups: quasi-planar, fully planar, and vertically-integrated arrays.

2.1.1 Quasi-Planar UWB Antenna Arrays

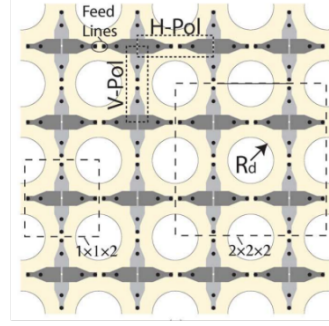
- Tightly-Coupled Dipole Array:*** Munk [29] for the first time proposed the idea of utilizing mutual coupling between the arrays to have a bandwidth of around one decade. It was then patented by Harris Corporation [30]. This antenna array configuration is widely known as a tightly-coupled dipole array (TCDA) in which mutual coupling between the dipole elements is utilized to achieve wide bandwidths. This technique is based on Wheeler's current sheet technique [31]. Such antennas have been reported to operate from 2-18 GHz [32] shown in Figure 2.1, and 8-40 GHz [33]. Aside from these benefits, these arrays can be complicated in design and fabrication since they involve complex feeding and multi-layer dielectric superstrate layers.

2.1.2 Fully-Planar UWB Antenna Arrays

- *Planar UWB Modular Antenna Array:*



(a)



(b)

Figure 2.2: Planar UWB modular antenna array covering 7-21 GHz (a) complete design view (b) feeding view [34].

Among the fully-planar array configurations, the planar UWB modular antenna (PUMA) array is based on the previously discussed Munk's implementation of the current sheet concept. PUMA arrays are fully-planar having multilayer PCB implementation and thus are low-cost and less complex than the vertically-integrated UWB arrays. These arrays have a bandwidth of around 5:1 [20, 34, 35] or 6:1 [36]. A depiction illustrating the geometric configuration of the PUMA array can be seen in Figure 2.2 [34].

- *Planar Tightly-Coupled Dipole Array:* The design presented in [37] and shown in Figure 2.3 is a full planar version that is based on the Planar Tightly-Coupled Dipole Array (PTCDA). The reported bandwidth is around 28 GHz covering 17.1-45.6 GHz. However, due to its integrated balun feeding mechanism, the design has the same complexity as the quasi-TCDA case but with comparatively lower cost due to the full planar PCB structure.

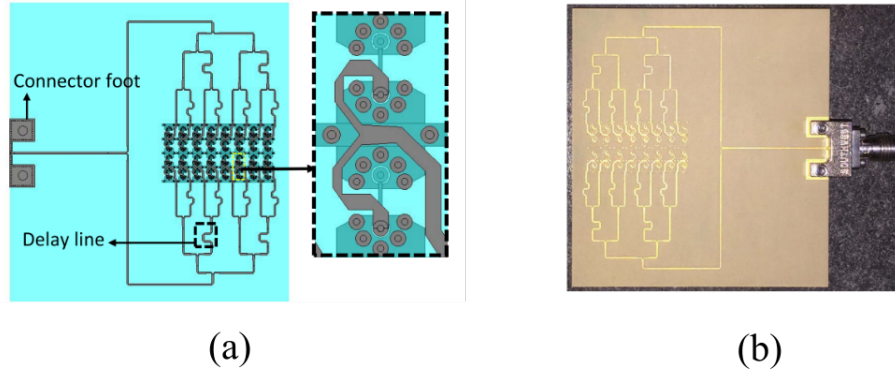


Figure 2.3: Planar UWB tightly-coupled array covering 17.1-45.6 GHz (a) feeding network for the array (b) feeding rear view [37].

- **Microstrip Patch Array:** Among the planar UWB antenna arrays, microstrip patch arrays have the lowest bandwidth but are less costly and complicated compared to other planar UWB arrays discussed so far. Although these antennas also have a multi-layer structure, they are considered fully planar due to their implementation on a simple PCB structure [38]. The stack-patch configuration has been reported to have a fractional bandwidth of around 20% [39, 40].

2.1.3 Vertically-Integrated UWB Antenna Arrays

- **Vivaldi or Tapered-Slot Array:** The other most famous category for UWB arrays is the tapered-slot array (TSA) [42–44] also known as the vivaldi array. These arrays can provide a wide bandwidth performance for up to a decade [45, 46]. For a single-polarization, TSA has a planar structure and is printed on PCB sheets. However, when dual-polarization is required, array elements are arranged in an egg-crate lattice and the whole structure is vertically oriented. Such dual-polarized TSA involves high complexity in the design and fabrication, needs an electrical connection between the ele-

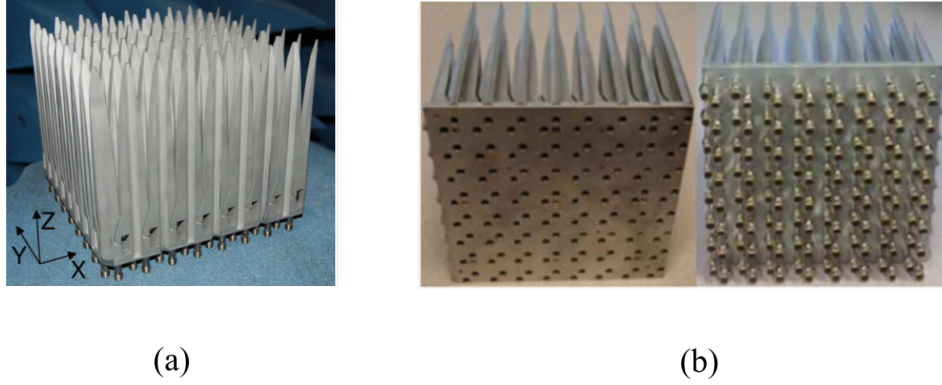


Figure 2.4: Tapered slot array covering 0.7-9 GHz (a) isometric (b) back view [41].

ments to avoid resonances, and also produces high cross-polarization levels in the diagonal plane [47–49]. In some cases of modular TSA designs, special machining is required (see Figure 2.4) and is difficult to implement above the X-band [41, 50–52].

- Dipole-Like UWB Array:*** Similar to TSA, these dipole-like arrays are also vertically integrated but can provide low cross-polarization compared to TSA and do not require an electrical connection between the antenna elements. They can be found in literature as Antipodal Vivaldi Antenna (AVA) [53] or Balanced Antipodal Vivaldi Antenna (BAVA) [54–57]. These antennas have wider bandwidth for up to a decade [56]. Similar to the dipole-like configuration, bunny ear arrays were introduced as a low-profile alternative to the TSA [58, 59]. These antennas have been reported to have an absolute bandwidth of up to 5:1 and its geometry can be seen in Figure 2.5 [58]. Similar to the TSA, vertical integration of the bunny ear array is complicated to implement for dual-polarization, especially above the Ku-band. This configuration has also some issues related to efficiency and power handling [60].

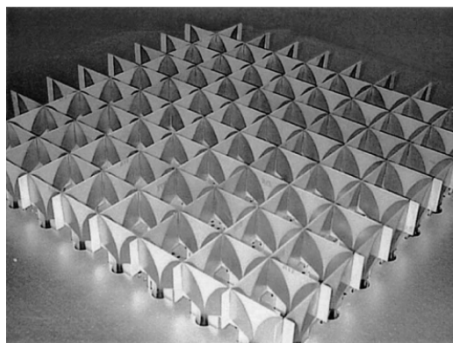


Figure 2.5: Low-profile bunny ear dual-polarized array covering 1-5 GHz [58].

2.2 Standalone UWB Antennas

Standalone UWB antennas have been widely reported in the literature for several decades and covering all of these antennas is itself a separate topic. Here, to provide context, a brief overview of standalone UWB antennas is presented. For a comprehensive study on planar UWB antennas, the reader is encouraged to consult several detailed books such as [8, 61–63]. The majority of the planar standalone UWB works have focused on the common commercial 3.1-10.6 GHz frequency band while the non-planar UWB antennas have covered frequency range from 2-32 GHz.

2.2.1 Planar UWB Antennas

- ***Frequency Independent Antennas:*** These Frequency Independent Antennas (FIAs) antennas are also known as self-complementary antennas. This category includes spiral, sinuous, log-periodic, biconical, and bow tie antennas [64]. The input impedance, gain, and circularly polarized radiation pattern of these antennas remain stable and consistent across wide bandwidths [64]. The geometry shown in Figure 2.6(a) has an absolute

bandwidth of 15 GHz [65].

- UWB Printed Monopole Antenna:*** Planar and printed monopole antennas are excellent choices for UWB wireless technology due to their extensive impedance bandwidth and nearly omnidirectional azimuthal radiation pattern. Numerous variations of planar monopole antennas, often referred to as planar discs, have been documented, resulting in wide bandwidths of up to 12.5 GHz covering the 1.1-13.5 GHz band [66–68]. Figure 2.6(b) shows an example geometry of a UWB printed monopole antenna [68] having an absolute bandwidth of around 7 GHz.

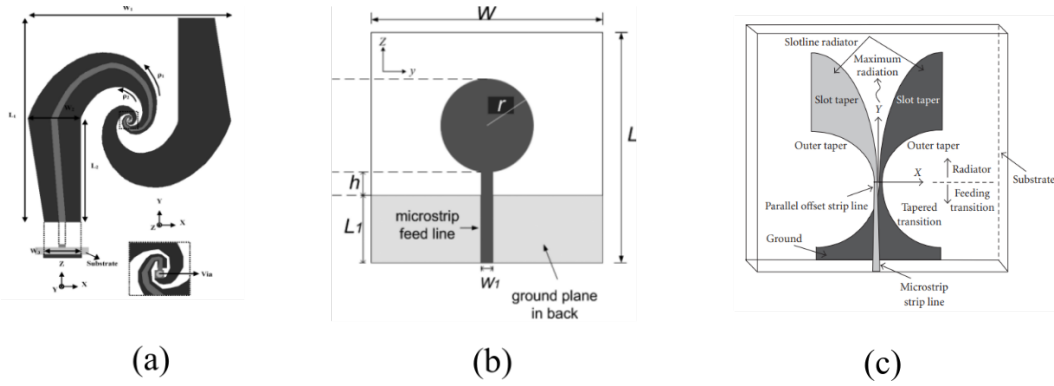


Figure 2.6: Geometry layouts of UWB planar antennas (a) a UWB spiral antenna having absolute bandwidth of 15 GHz [65] (b) a UWB printed monopole antenna having absolute bandwidth of 7 GHz [68] (c) a UWB tapered slot (vivaldi antenna) having a bandwidth of 17 GHz [69].

- UWB Tapered Slot Antenna:*** Tapered Slot Antenna [70] belong to the category of end-fire radiating antennas, originating from vivaldi aeriels [71]. The majority of these antennas have a bandwidth of around 17 GHz (1-18 GHz) [69] (shown in Figure 2.6)(c) or as high as 42 GHz (9-50 GHz) [72] and are widely utilized for radar and ground-penetrating radar applications. However, these antennas suffer from high cross-polarization levels of

usually greater than -15 dB and also get complicated in terms of geometry configuration (vertically integrated) when dual-polarization is required.

2.2.2 Non-Planar UWB Antennas

- **Helical Antennas:** These antennas were first introduced for space communication by Kraus [73]. A few variants of the helical antenna were later presented to broaden the bandwidth up to 2 GHz [74]. These antennas come with a single circular polarization whereas, for weather radars, dual linear polarization is required.

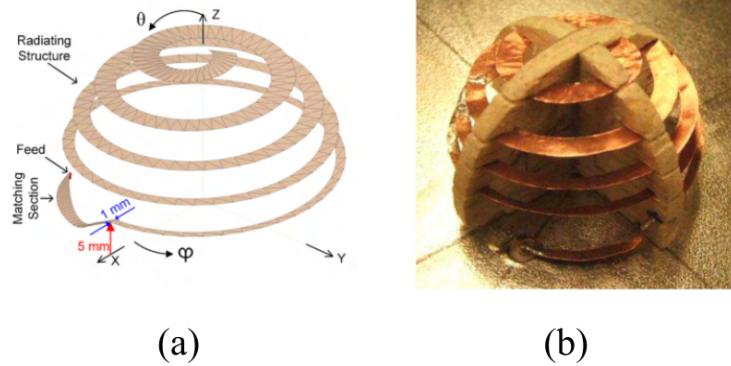


Figure 2.7: A helical antenna having a bandwidth of 2 GHz (a) geometry model (b) fabricated prototype [74].

- **Quad-Ridged Horn Antenna:** Traditional horn antennas are famous for their distinctive features: ease of feeding, relatively straightforward construction, high gain, and outstanding peak power handling capability. However, the advancement of wideband testing systems has constrained the use of these conventional horn antennas. Acting as band-pass filters, horn antennas exhibit limited bandwidth, thus prompting a pressing requirement for techniques to broaden the antenna operational frequency range. *The QRHA represents the horn category with the broadest bandwidth. In these antennas,*

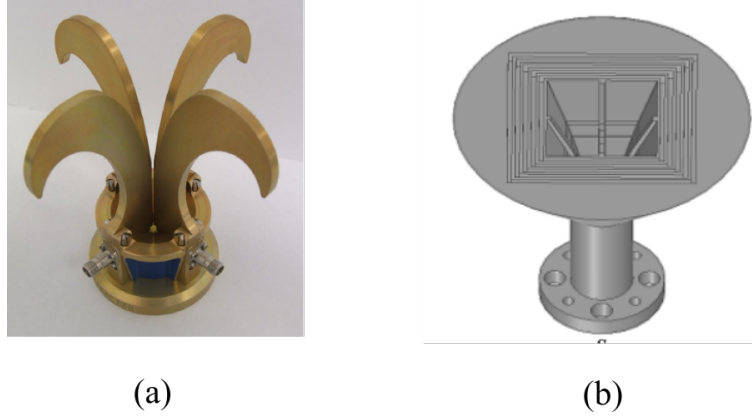


Figure 2.8: QRHA geometries (a) CNC-machined MVG commercial QH2000 antenna covering 2-32 GHz [18] (b) direct metal laser sintering (DMLS) fabricated QRHA covering 45-110 GHz [79].

waveguides are loaded with dual- or quad-ridged structures thus lowering the cut-off frequency of the dominant mode and increasing the bandwidth performance. For the detailed theory and discussion on ridged waveguides, the reader is encouraged to consider these pioneering works [2, 75–78]. The majority of the dual- or quad-ridged works have targeted the 1-18 GHz or 2-20 GHz frequency band thus reporting maximum absolute bandwidth of up to 20 GHz. Few commercially available QRHAs have covered frequencies from 2-32 GHz [18] providing an absolute bandwidth of 30 GHz. QRHA in [79] has shown extreme broadband performance having an absolute bandwidth of 65 GHz covering from 45-110 GHz.

2.3 Literature Review: Summary and Remarks

Figure 2.9 shows the comparison of both UWB antenna arrays as well as UWB standalone antennas based on their absolute bandwidth performance vs. cost or design complexity. From the detailed literature review, it can be noticed that in the case of UWB antenna arrays, although microstrip patch arrays have less cost

and complexity, however, this comes at the expense of the lowest bandwidth performance. TCDAAs have been shown to provide the largest bandwidth performance but at the expense of cost and complexity in the design and feeding mechanism as well as the use of multiple superstrate layers. In the case of standalone UWB antennas, the planar PCB-based UWB antennas have less associated cost and complexity, but they can provide a maximum bandwidth of up to 20 GHz. Moreover, it is difficult to implement dual-polarization in many of these cases except QRHA. Vivaldi or TSAs have shown wideband performance up to 40 GHz but their geometries become complicated when they are vertically integrated for dual-polarization and their fabrication is thus limited to the Ku frequency band. Also, these antennas have been shown to have high cross-polarization levels of around -12 dB.

Figure 2.10 shows another view of the comparison between arrays and standalone antennas for their absolute bandwidth and HPBW. It is worth noting that the previously used 'Stop-Sign' microstrip array meets the beamwidth requirements but has a very narrow absolute bandwidth and a complex multi-layer geometry, including a feeding network. On the other hand, commercial standalone QRHAs offer a wider bandwidth but suffer from wide beamwidth performance, making them unsuitable for UAV-based measurements. As illustrated in the figure, the first goal of this research is to adapt these commercial QRHAs for UAV-based radar characterization by reducing their beamwidth performance through the integration of dielectric lenses. The second goal is to propose a new UWB QRHA that meets both bandwidth and HPBW requirements.

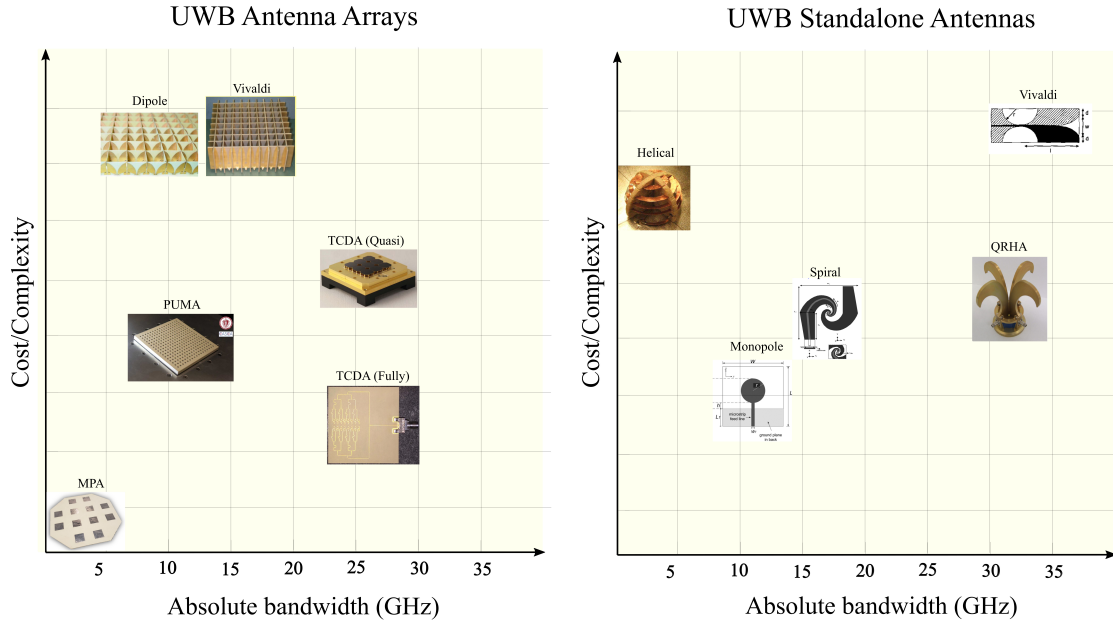


Figure 2.9: A comparison diagram of the UWB literature works based on their absolute bandwidth vs. cost/complexity (to have a fair comparison in terms of absolute bandwidth, only those designs are considered having a lower frequency ($f_L < 10$ GHz)).

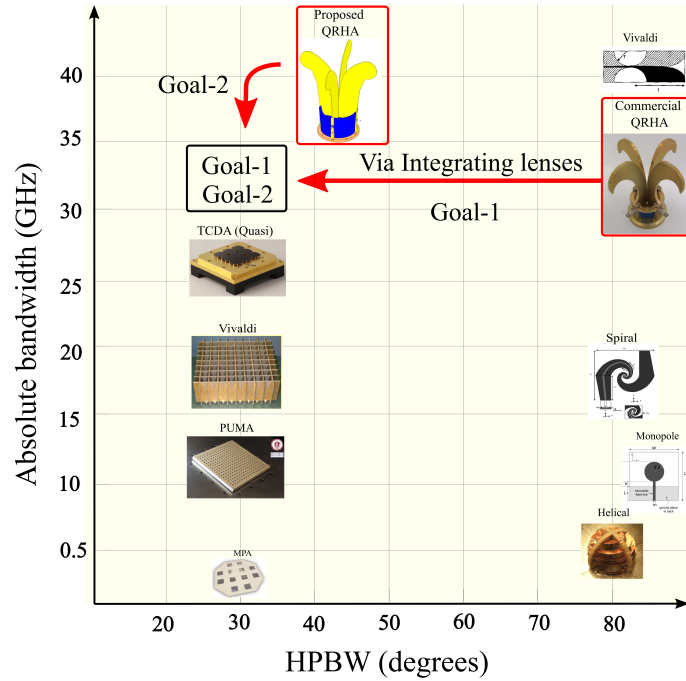


Figure 2.10: A comparison diagram of the UWB literature works based on their absolute bandwidth vs. HPBW.

Based on the literature works and discussions, it can be concluded that QRHAs have the potential to provide wide bandwidth performance (can have absolute bandwidth of around 30 GHz) without much increase in the cost or complexity of the design compared to the UWB arrays. QRHAs can be fabricated using modern 3D-printing, direct metal laser sintering (DMLS), and CNC-machining. Moreover, these antennas also provide the ease in implementation of dual-polarization and have a simple feeding mechanism compared to other available options. Finally, it will be shown in the coming chapters that QRHAs are favored over UWB arrays for UAV-based radar characterization. They offer a broader absolute bandwidth than arrays and maintain similar narrow beamwidth performance while using a comparatively simpler geometry and feeding mechanism.

2.4 Phased Array Radar: A Quick Overview

In the recent decade, there have been significant efforts to transition from conventional mechanically steered dish-based radar systems such as WSR-88D and NEXRAD to fully digital dual-polarized phased array radars equipped with electronic scanning capabilities [80]. Traditional weather and surveillance radars rely on the mechanical movement of the radar dish to scan different areas of the atmosphere. On the other hand, phased array radar (PAR) presents a groundbreaking innovation through its stationary flat panel antenna comprising a fixed grid of individual elements, each with the ability to transmit and receive signals independently. This radar technology allows for electronic steering of the radar beam, providing users with precise control over the scanning process, including timing, location, and direction. Moreover, this technology allows the radar to focus its

beam exclusively on areas where storms or desired targets are present, avoiding unnecessary scanning of clear-air regions thus enabling faster updates. Within a minute, the entire sky can be scanned, revealing critical weather phenomena such as deadly storms, hail, or tornadoes. Additionally, it offers vital surveillance support for homeland security and military applications [81].

Dual-polarization capability is required to accurately estimate and identify weather phenomena. Dual-polarization is the ability of the radar to simultaneously transmit and simultaneously receive (STSR), alternatively transmit and simultaneously receive (ATSR), and alternatively transmit and alternatively receive (ATAR) electromagnetic waves in horizontal and vertical orientations. This allows for gaining more as well as accurate information about the characteristics of the target. Some of the key benefits of a dual-polarized weather radar compared to a single-polarized radar are proper identification of precipitation (rain, snow, hail, sleet), rain-fall estimation, hail and debris detection, and hydrometeor classification. To achieve these benefits, the dual-polarized radar system has to satisfy specific requirements in terms of correlation coefficient (ρ_{HV}), differential reflectivity (Z_{DR}), and specific differential phase (K_{DP}) [82]. Errors in the measurements of ρ_{HV} due to the high cross-polarization or radiation pattern mismatch, can significantly affect the ability of a radar to accurately distinguish between different types of hydrometeors (precipitation particles), while rainfall rate estimates can be significantly affected by biases in Z_{DR} and K_{DP} due to high cross-polarization and mismatch between the radiation patterns of the horizontal (H) and vertical (V) polarizations [83–86]. These parameters thus enforce stringent requirements on the mismatch between the co-polarization antenna patterns of no more than 0.1 dB and cross-polarization level of less than -45 dB for radars operating in STSR mode [87]. This entails the use of precise antennas both in radars and as a probe

for calibration as well as accurate in-situ calibration measurement techniques.

- ***Calibration or Measurement:*** For accurate evaluation of the radiation performance of the antenna or antenna array, specialized indoor or outdoor measurement facilities are required. Such facilities require specific RF equipment, procedures, and physical space (in the case of non-mobile radars) to conduct precise measurements. However, building such facilities has space and cost constraints at the antenna installation site. Moreover, the indoor performance of the antenna/array can deviate from the outdoor measurements due to several external factors such as rain, temperature, humidity, dust, wind, RF system interference, ground clutter, and component failure. To ensure the adequate performance of a radar or communication system in a real operational environment, it is crucial to conduct an in-situ (real-time) characterization of the employed antenna. This involves evaluating the antenna’s performance in its actual operating conditions and environment [14, 88–90].

Several outdoor in-situ antenna measurement techniques have been used in the past using tethered balloons [91], heliborne setup [92], and measuring the near-field using telescope [93]. These techniques are costly, demand labor, and are time-consuming. A recent surge in UAVs, has garnered significant attraction for their use in antenna measurement and radar characterization. Several works using UAVs for this purpose can be found in the literature [94–109]. UAV-based radar characterization is a cost-effective and portable solution. Moreover, using UAVs, the characterization of the radars can be done in cylindrical or even spherical modes, which can provide extra freedom and useful information which is not usually possible in case of the stationary measurement setups. Furthermore, this technique is capable of removing the unwanted ground noise/clutter

by flying the drone in the vicinity of the radar compared to the present expensive stationary mounted calibration system where the ground clutter adversely affects the measurement data which can lead to false detection and forecasting. With all the above-mentioned benefits associated with using UAVs for radar characterization, there are some challenges associated with it. Other than the UAV's flight performance and stability concerns, special attention is required to RF performance when choosing probe antennas to be mounted on the UAV. These probe antennas have to satisfy certain strict requirements set for weather radar calibration in terms of half-power beamwidth (HPBW), and co-polar mismatch between the patterns. The interference of the electromagnetic energy between the probe antenna and the UAV structure can lead to unwanted scattering and diffraction effects, thus producing undesired ripples and mismatch in the co-polar patterns as well as degrading the polarization performance. As a consequence, the overall performance of the system can be degraded.

The authors in [15–17] have recently conducted a detailed study on the selection of the probe for UAVs and have compared the flight vs RF performance for different antenna probes specifically using dipole, single patch antenna, patch antenna array, and horn antennas. These works have concluded that using dipole and microstrip patch antennas having non-directional or omnidirectional radiation patterns (having $\text{HPBW} \geq 90^\circ$), significantly interfered with the UAV structure and hence degraded the RF performance. However, due to their lightweight structure, these printed antennas exhibited good flight performance. In contrast, the microstrip patch antenna array ($\text{HPBW} = 28^\circ$) and the horn antenna ($\text{HPBW} = 41^\circ$) showed excellent RF radiation performance due to their narrow HPBW performance compared to dipole and single patch

antenna. Among these two choices, although the horn antenna showed excellent RF performance, the patch antenna array was chosen due to its lightweight and compact structure compared to the bulky horn antenna.

This study has also set some requirements for the RF performance of the probe to be mounted on the UAV for in-situ measurements. The desired probe must have a 3 dB HPBW of around 40° , an H- and V-co-polarization mismatch of no greater than 0.1 dB, and a ripples magnitude of less than 0.15 dB, within $\pm 45^\circ$ field of view.

2.5 Summary

This chapter has provided a comprehensive literature review on UWB arrays and standalone antennas. Various comparisons between these two categories were presented and analyzed. It was emphasized that for UAV-based radar characterization, beam steering is not required. Therefore, standalone UWB antennas are favored over complex arrays as long as they can meet both bandwidth and beamwidth requirements. Standalone QRHAs were identified as suitable candidates for UAV-based radar characterization, capable of delivering UWB performance. Their beamwidth can be reduced to meet the requirements through the integration of dielectric lenses, which will be discussed in the upcoming chapters. Additionally, this chapter has given an overview of PAR in-situ calibration techniques, including their RF requirements. In the next chapter, commercial QRHAs are tested for their beamwidth performance based on the far-field measurements when these probes are installed on the UAV platform.

Chapter 3

Commercial Ridged Horn Antennas: Testing and Validation

3.1 Introduction

In the preceding chapter, several literature studies were presented related to using UAVs for in-situ characterization of radars. Through various UAV-based far-field antenna measurements as well as EM simulations, these works have established detailed guidelines and have studied the effects of coupling between the UAV structure and the probe antenna in order to mitigate many error sources as well as provide a selection criterion to choose the best type of probe antenna that meets the mission requirements.

It was discussed that in previous UAV-based works, a very narrowband microstrip SS patch array having a bandwidth of only 70 MHz was used for the characterization of radars. The main disadvantage of using narrow-band probes for a particular application is the dedication of a UAV platform and probe for a specific mission. Assigning a UAV platform for a single task is not cost-effective. On the other hand, reusing the same UAV platform for other bands requires tedious characterization and RF calibration of the probe with the UAV platform to guarantee a good antenna test during flight. Ideally, a single probe mounted on a UAV that operates in a wide frequency bandwidth is desirable.

Through a detailed literature review, It was identified that a QRHA is well-suited to deliver exceptional UWB performance, potentially serving as a “one-size-fits-all” antenna capable of encompassing the majority of radar and communication system bands operating within the 2-32 GHz range. The advantages of ridged horn antennas were discussed, along with their ability to provide extensive bandwidth and ease of dual-polarization, at a low cost with reduced complexity compared to arrays or other standalone antenna options. Nonetheless, amidst the various challenges discussed before, one notable concern pertains to their wide beamwidth performance, particularly in the lower frequency range between 1-6 GHz, as well as beamwidth variability throughout the entire band. It was claimed that the wide beamwidth performance of the commercially available dual-ridged horn antenna (DRHA) or QRHA, limits their use in several applications including UAV-based in-situ characterization of radars. This chapter thus validates this claim and presents the analysis based on the measurements conducted using these commercially available ridged horn probe antennas when mounted on a UAV platform.

3.2 UAV Metrology System Description

Nowadays, antenna test and measurement, and radar characterization using UAVs are very popular, ranging from micro and small to medium and large frames. It has seen a fast-paced evolution in the past decade, in virtue of the increased availability of commercial off-the-shelf flight solution suites with a high degree of precision and performance at lower costs. The proposed metrology system, shown in Figure 3.1, was developed to provide a UAV platform with a single front-end equipment to perform in-situ antenna patterns, calibration of communication,

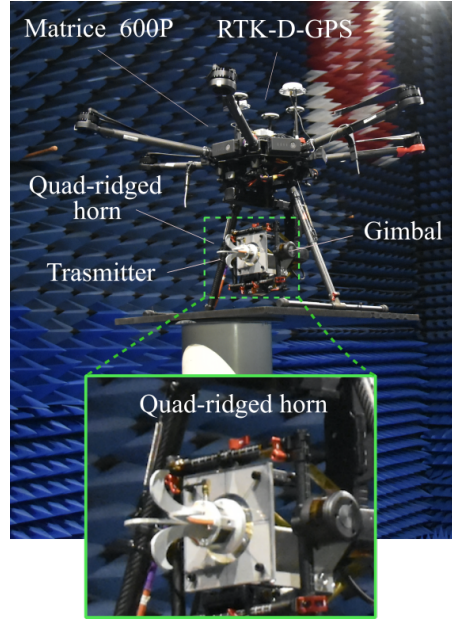


Figure 3.1: Proposed UWB (2-32 GHz) UAV-based metrology system using [18] for in-situ testing of antennas in radars and communication systems.

and radar systems that operate in the frequency range from 2 GHz to 32 GHz using commercial DRHA (SH2000 [110]) and QRHA (QH2000 [18]). The system is mainly composed of three subsystems. The UAV platform, the gimbal, and the front-end subsystem which is composed of a UWB dual-polarized antenna probe interconnected with a CW transmitter source. The summary of the system specifications is listed in Table 3.1.

3.2.1 UAV platform

The UAV platform is the hexacopter DJI Matrice 600P that provides a maximum flight time from 20 min to 60 min with three sets of interchangeable batteries. This platform provides excellent stability for a maximum payload of 2 kg. The system was tested with a wind load of up to 20 mph providing excellent test results. Real-time kinematic differential GPS (RTK D-GPS) is integrated with the UAV platform, and it provides position accuracy lower than 2 cm.

Table 3.1: UAV metrology system specifications.

Category	Specifications	Value
UAV Platform	Model/Maker Dimensions Weight (no payload) Max. takeoff weight Position accuracy Position accuracy Max. angular velocity Max. pitch angle Max. ascent speed Max. descent speed Max. serv. ceiling Hovering time Operating temp.	Matrice 600P/DJI 1.66 m x 1.52 m x 0.727 m 10 kg 15.5 kg GPS: ± 5.5 m to ± 0.5 m RTK: ± 1 cm to ± 2.0 cm Pitch: $300^\circ/\text{s}$, Yaw: $100^\circ/\text{s}$ $25^\circ/\text{s}$ 5 m/s 3 m/s 2500 m 20 min (up 1 hr with 3 sets of batt.) -10°C to 40°C
Gimbal	Model/Maker Operation modes Dimensions MIU Connectivity Operation. Freq. Running time DC voltage Operating temp.	Ronin-MX/DJI Free, Follow, FPV 28 cm x 34 cm Independent Bluetooth and USB connections 2.4 GHz 3 hrs 12 VDC -15°C to 50°C
Transmitter	Model/Maker Freq. operation Tx. power Dynamic range Amplitude resol. Phase resol. Dimensions DC voltage Weight	Windfreak technologies 10 MHz to 32 GHz 20 dBm (CW) 50 dB 0.01 dB 0.01° 4 cm x 4 cm x 2 cm 12 VDC < 0.1 lb

3.2.2 Gimbal

The DJI Ronin-MX gimbal, which is used to carry the antenna probe, is very stable and flexible for accurate position alignment between the AUT and the probe. Independent IMUs make this gimbal to be accurate. This gimbal can be controlled automatically and manually or by point-of-interest (POI) through the mission planner application, with an accuracy of 0.02° .

3.2.3 Transmitter

On the gimbal, a CW transmitter source and antenna are used for the AUT test in receive mode. The transmitter is a two-channel synthesizer that generates signals from 10 MHz to 14 GHz. An active multiplier is used to expand the frequency up to 34 GHz with an output power up to 30 dBm. High dynamic range and amplitude (0.001 dB), and phase resolution (0.01°), make this transmitter an ideal candidate for this UAV metrology concept.

3.3 Probe Antenna and UAV Considerations

In the proposed system, two commercially available UWB antenna probes: SH2000 and QH2000 are mounted on the drone to perform radiation pattern measurements in the E- and H-planes. Figure 3.2 shows the geometry models of the two commercial probes used, realized gain, and E-plane and H-plane HPBW comparison. These figures are generated from the data provided in the data sheets in [18, 110]. Table 3.2 summarizes the overall performance of these two antenna probes.

In many cases, commercial probes, mounted on UAV platforms, are used with the basic assumption that the probe antenna performance will be not affected by

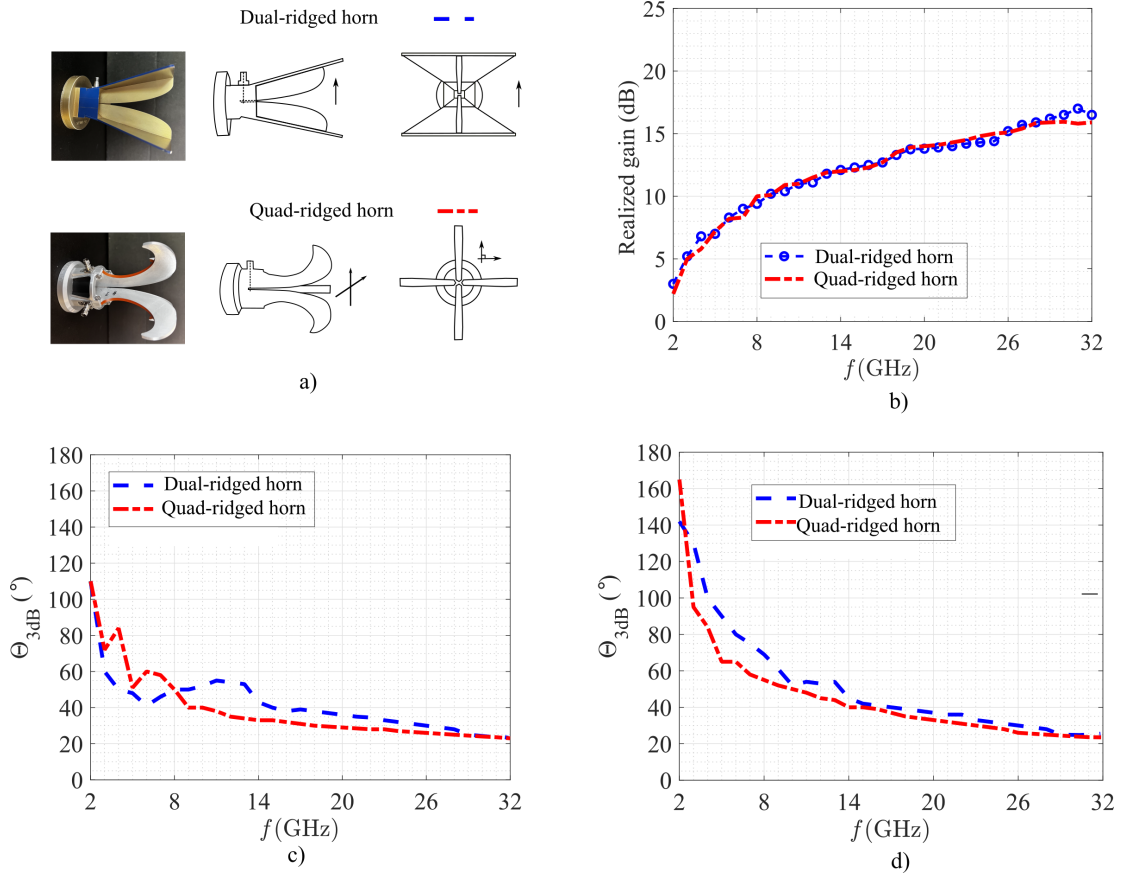


Figure 3.2: Commercial antenna probes used in this study. (a) pictures and basic geometries of the DRHA (SH2000) and QRHA (QH2000) (b) realized gain for both antenna probes (c) E-plane HPBW of both antenna probes using V-port (d) H-plane HPBW of both antenna probes using V-port.

the electromagnetic scattering and coupling of the energy between the probe and the UAV platform. This assumption can be valid when the electrical dimensions of the antenna are large in comparison with the UAV platform. For microwave or mm-Wave applications, the UAV platform is always larger than the electrical dimension of the antenna, thus the characterization of the probe with a UAV platform is crucial for accurate measurements of the radiation patterns of the antenna under test (AUT). Previous work done by the authors in [15, 16, 108] performed exhaustive numerical simulation analysis to characterize the impact of

Table 3.2: Summary of performance comparison of the commercial probes.

	Freq. (GHz)	HPBW ($^{\circ}$)		Gain (dB)	RL (dB)	Size (in)	Weight (lb)
Probe Type	(f_{min}/f_{max})	E-plane (f_{min}/f_{max})	H-plane (f_{min}/f_{max})	(f_{min}/f_{max})		(x-y-z)	
Dual-ridged	2/32	110/22	142/23	3/16.5	> 7	4.13 x 2.4 x 4.1	1.1
Quad-ridged	2/32	110/23	165/23.5	2.2/16	> 6	4.32 x 4.13 x 4.13	0.53

electromagnetic scattering of the drone on the radiation performance of the probe antenna. These studies have established a set of requirements in which the 3 dB HPBW of the antenna should be around 40° when using Matrice 800 and Matrice 600 platforms, in order to have less scattering between the antenna and the UAV [15].

Another important aspect of guaranteeing an antenna or radar test using a UAV platform relies on flight stability, position accuracy, dynamic range, and alignment between the AUT and the probe. Matrice 800/600 drone uses mature and proven technology that enables stable test operation in wind conditions up to 20 mph with a flight time of 20 min to 60 min using 1 to 3 sets of batteries. Differential GPS based on a real-time Kinematic (D-GPS RTK) algorithm was tested on flight conditions. Previous results demonstrated a drone position with an accuracy ± 2 cm (in x-,y- and z-axis) [108, 109]. Having an accurate position of the drone during each mission is quite important to compensate for any position deviation due to the wind of anomaly in the flight path. A high dynamic range is required for antenna patterns test, about 30 dB SNR is desirable for far-field antenna patterns. In the proposed system, the synthesizer source can provide a maximum power of 30 dBm to compensate for the path loss, especially at high frequencies, a gain of 30 to 40 dB is required in the receiver path. The probe and transmitters were mounted on a Ronin-MX, and this was implemented with a smooth-track algorithm that ensures smooth movements during flight conditions.

Probe, synthesizer, and battery position on the gimbal are critical to guarantee balance and stability with a low power consumption in the motors and IMUs. The slip ring on the gimbal allows 360° continuous rotation in azimuth. This feature enables the possibility of a wide range of probe position selection with respect to the UAV platform that helps to minimize the induced scattering of the UAV when the probe is transmitting.

3.4 The Need for Narrow Beamwidth at Low Frequencies: Preliminary Results

For measurements, both SH2000 and QH2000 are used as a probe and also as an AUT. Since both probes exhibit very similar performance (see Table 3.2), measurement results for cases using a single-polarized SH2000 probe are presented for the sake of simplicity. Full characterization of the probe with and without the UAV was performed. The indoor far-field anechoic chamber of the Advanced Radar Research Center (ARRC) was used. For this study, a frequency range from 3 GHz to 12 GHz is used due to the limitations of the antenna range system of the chamber. Figure 3.3 illustrates the measurement plan and configuration used to achieve both co-and cross-polarization patterns with and without the drone cases.

For illustration, measured radiation patterns at the lower and the upper-frequency bands are shown in both E and H planes and for both polarizations: vertical polarization (V-polarized) and horizontal polarization (H-polarized). These results are shown in Figures 3.4 and 3.5. The solid lines represent the co-polar patterns, and the dashed lines represent the cross-polar patterns. The black color represents the AUT measured with an isolated probe (no drone). The solid blue and red lines represent the patterns of the AUT measured with a probe mounted

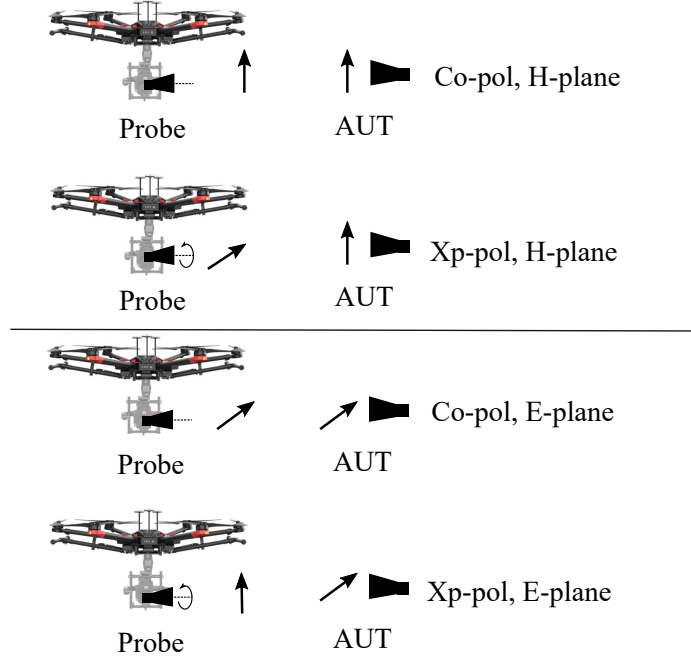


Figure 3.3: Illustration of the far-field chamber measurement setup used for characterization of the AUT using the drone.

on the drone.

From these figures, it can be seen that when the antenna is V-polarized (Figure 3.4), the boresight cross-polarization for the drone case is less than -25 dB at low frequencies in both E and H planes. However, when the antenna is H-polarized as shown in Figure 3.5, the cross-polarization for the drone case gets worse at low frequencies, especially in the H-plane where it is around -16 dB. This is due to the fact that at low frequencies, the beamwidth is wider and the fields strongly interact with the drone structure and create ripples in the co-polar patterns as well as increasing the cross-polarization contamination. On the other hand, at higher frequencies, for both polarizations, the cross-polarization levels are not significantly affected because the beamwidth at higher frequencies is narrower in both planes and thus the scattering from the UAV platform is minimum. Moreover, it can also be noticed that in both of these polarizations, there is also a mismatch/asymmetry

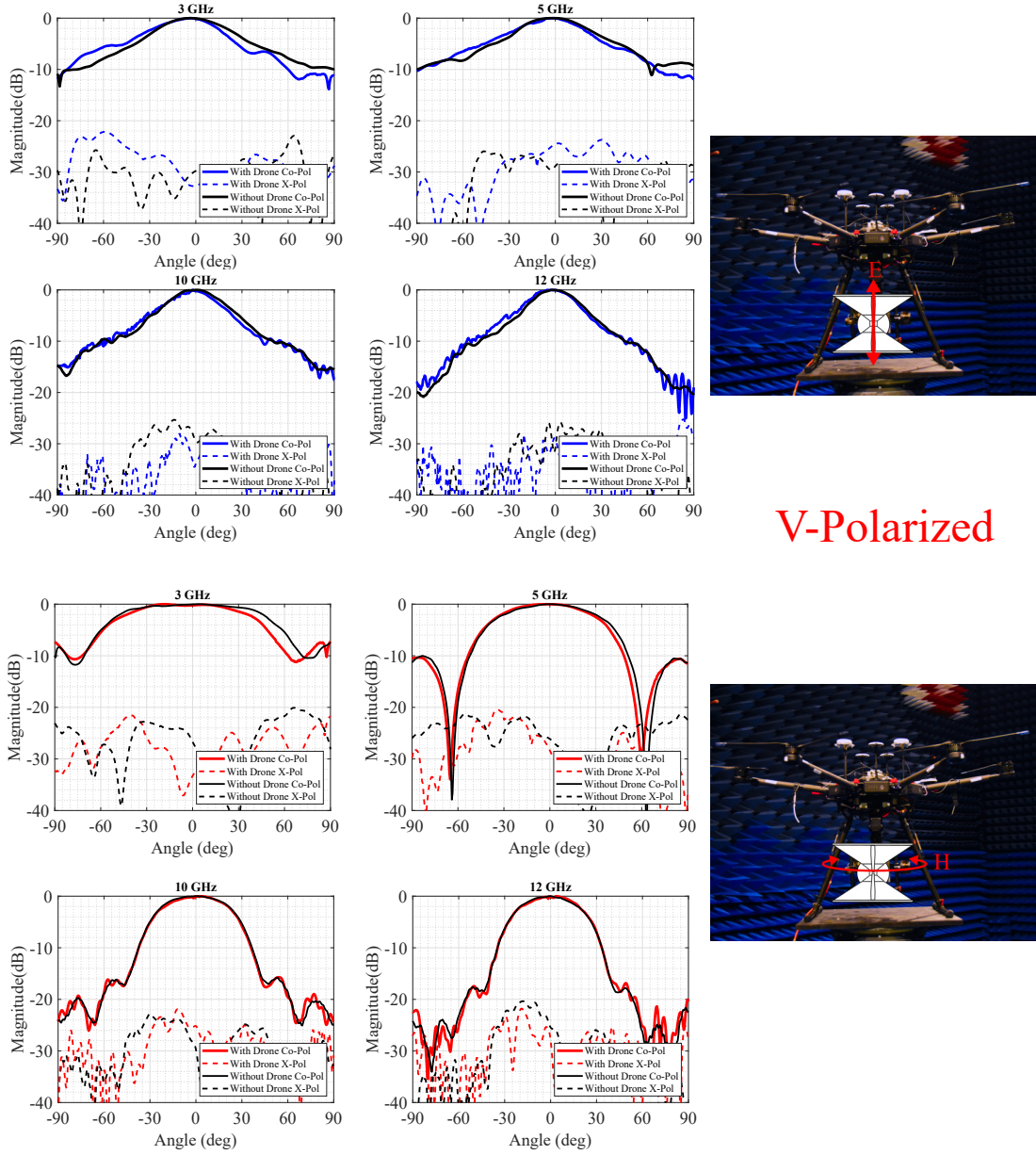
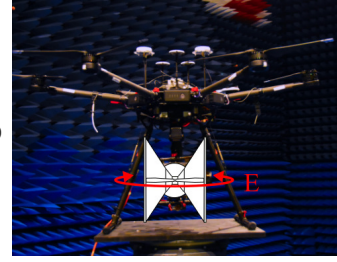
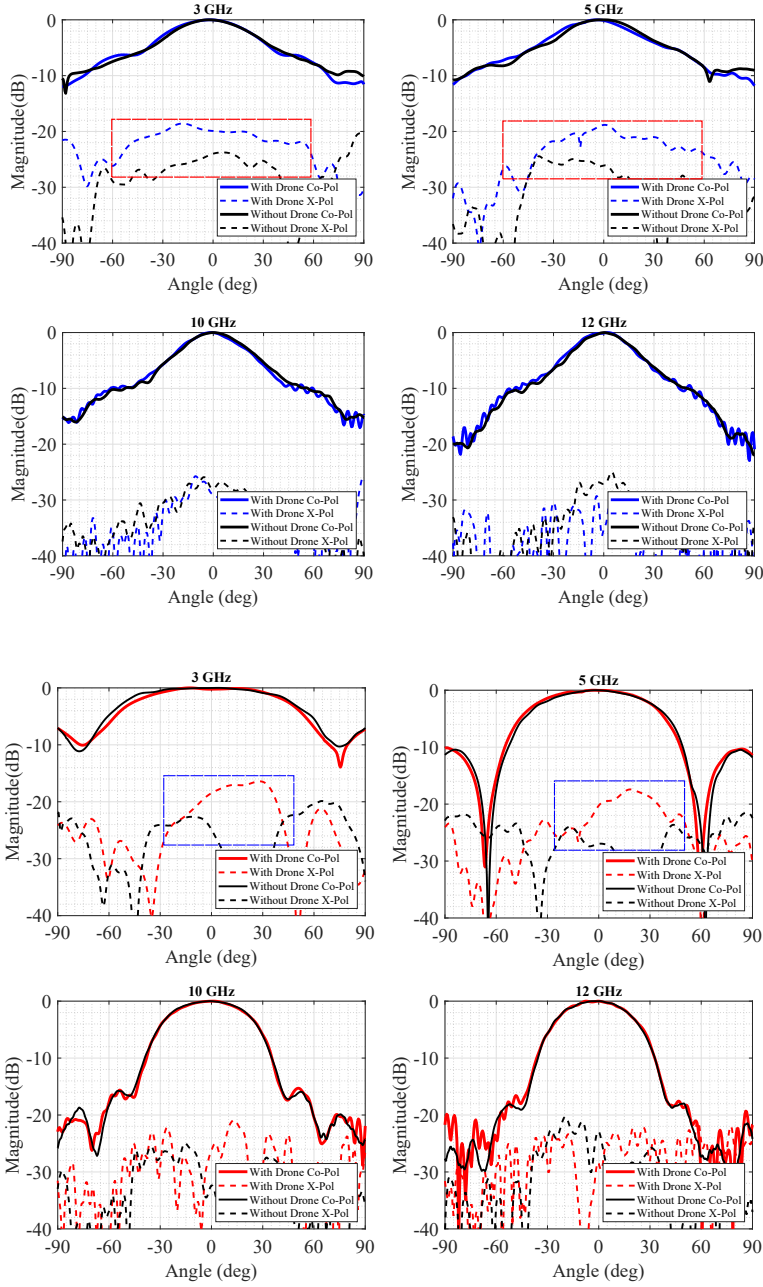


Figure 3.4: Comparison of the measured radiation patterns in E and H planes at different frequencies for with and without drone cases, when the antenna is vertically polarized (V-polarized) .

in the co-polar patterns of the E- and H-planes when the UAV is taken into consideration. These results validate the claim raised and also necessitate the need for beamwidth reduction in these probes for in-situ characterization of radars using UAVs.



H-Polarized
90° rotated



Figure 3.5: Comparison of the measured radiation patterns in E and H planes at different frequencies for with and without drone cases, when the antenna is horizontally polarized (H-polarized).

3.5 Summary

Ridged horn antennas are useful for UAV-based measurements since they offer wide bandwidth and compact size. A single wideband ridged horn probe can be used for the characterization of radars in the S-, C-, X-, and Ku-bands, instead of using multiple probes for characterization and measurements. This is cost-effective and time-saving. However, on the dark side, these ridged horn antennas suffer from wider beamwidth performance, especially at low frequencies between 1-6 GHz. Through detailed measurement analyses, it is shown that specifically at low frequencies, when the beamwidth is wider ($> 40^\circ$), the fields strongly interact with the drone platform, thus increasing the cross-polarization levels by around 9 dB as well as causing a mismatch or asymmetry in all the co-polar patterns. These results indicate a requirement for reducing HPBW beamwidth at lower frequencies and maintaining consistent beamwidth across the entire band to effectively utilize these antenna probes for UAV-based characterization of radars. Several solutions to this problem will be discussed in detail in the forthcoming chapters.

Chapter 4

Theory and Design of Artificial Dielectric Material for Microwave and mm-Wave Applications

4.1 Introduction

This chapter presents in detail, the theory, design, and synthesis of a low loss tangent ($\tan\delta$) artificial dielectric material (ADM) with desired effective permittivity or effective dielectric constant (ε_{eff}) values for microwave and mm-Wave applications. A detailed review and comprehensive error analysis of the most common particle mixing techniques are conducted to guarantee an accurate design for high-performance ADM. Several analyses based on the geometries of various particles are carried out to identify the most accurate mixing model used in the design of the proposed substrate. Accurate analytical particle mixing models are highlighted for various particles. The proposed ADM is then designed and characterized based on the well-known two-phase Maxwell Garnett mixing theory, where the ε_{eff} of the ADM is synthesized depending on the geometry and the permittivity of mixing particles and the permittivity of the host material. The effects of the direction of excitation as well as the polarization of the incident field on the ε_{eff} of the anisotropic particle are analyzed and discussed in detail. The analysis is validated by fabricating and testing the proposed ADM in the microwave fre-

quency (S-band) at 3.5 GHz, as well as in the mm-Wave frequency (W-band) at 77 GHz.

The proposed method enables the use of existing high-performance materials that do not necessarily provide a low dielectric constant (ε_r) and low $\tan(\delta)$. For mm-Wave antenna applications, materials with a ε_r of 2 – 4, and $\tan(\delta)$ of less than 0.002 are desirable to maximize gain and radiation efficiency. Commercial RF substrates can satisfy those requirements, however limited thermal expansion coefficient and lamination difficulties increase the cost significantly. The proposed method enables the use of inexpensive materials that provide excellent thermal properties and great compatibility with a multi-layer fabrication process with desirable ε_{eff} and $\tan(\delta)$. Measured results show a reduction of 45% in the ε_{eff} and 38% in the $\tan(\delta)$ values in the S-band, and 32% and 72% reduction in ε_{eff} and $\tan\delta$, respectively, in the mm-Wave frequency band. The simulated, calculated, and measured results show a very good agreement.

4.2 Background

A complete understanding of the propagation of electromagnetic waves inside the dielectric material is of great importance in the modern applications of material design, remote sensing, aerospace, lens manufacturing, electromagnetic absorbers, carbon nanotubes, and polymers, etc. [111–113]. The dielectric properties of the material depend on the internal structure of the particle, its shape, and the fractional volume (f). Several analytical models or mixing rules exist in the literature that provide information about the dielectric properties via simplified modeling of the complex fields generated in or between the mixing particles inside the host material.

Material mixing rules are a set of algebraic formulations intended to calculate the ε_{eff} of the inclusions or particles inside the mixture based on the individual permittivities of the particles as well as their fractional volumes. These mixing rules differ based on several factors such as the structure or geometry of the particle (sphere, elliptical, disc, cubic, cylindrical, rod, needle-like, or any random shape) and their distribution inside the host material (aligned or randomly distributed). The geometry of the particles inside the host material determines the isotropic or anisotropic properties of the material. For instance, a spherical inclusion is considered isotropic as the ε_{eff} evaluated in x -, y -, and z -directions remains the same. This is because, the sphere is symmetric due to which the induced electric field inside the particle is uniform, thus making the polarizability (α) as a scalar quantity. On the other hand, in the case of anisotropic particles, ε_{eff} is a tensor and is evaluated in each direction of the applied electric field. Generally, for antennas or other microwave devices, ε_{eff} is evaluated in the z -direction. However, characterization of the material for its ε_{eff} in the x - or y -directions can be important in applications such as leaky wave antennas or in the devices where the surface waves are critical to the performance of the system.

Several works on controlling the permittivity of the material are presented for various applications such as multi-band arrays and integrated systems [114–119]. The concept of using perforations to control the ε_{eff} for the substrate integrated image guide (SIIG) was first reported in [114, 115]. Cylindrical metallic particles were embedded in the host medium in [117] to change the overall permittivity of the substrate material and to obtain frequency tuning for a microstrip patch antenna. In [118], linearly tapered slot antennas were presented for 30 and 94 GHz bands on the synthesized low-permittivity of the perforated substrate. A leaky-wave antenna based on a dielectric layer with periodic perforations covering

96 – 108 GHz frequency band was presented in [119].

Choosing the right printed circuit board (PCB) material is significantly important for both microwave and mm-Wave frequency bands. The main losses associated with PCB materials are radiation losses, dielectric, conductor, and surface-wave losses. None of these losses can be ignored at mm-Wave frequencies, specifically, the conductor losses, which become significant at mm-Wave frequencies. To achieve optimum performance from a PCB at any frequency of operation, several factors are evaluated: ε_r , $\tan(\delta)$, copper roughness, coefficient of thermal expansion (CTE), moisture absorption, and material thickness [120]. However, at high frequencies, some of these factors become more critical such as ε_r , $\tan\delta$, copper roughness, and material thickness [121].

At microwave frequencies, high ε_r material (8 – 500) can be used to achieve miniaturization of the circuit [122]. However, at mm-Wave frequencies, low ε_r material (2 – 4) is preferred in order to reduce losses in the material ($\tan\delta$). The conductor losses, which are either due to the material's surface finishing or copper roughness, become critical at mm-wave frequencies. The losses from the copper roughness are directly related to the frequency of operation through the skin depth. At higher frequencies, these losses are significant because the value of the skin depth becomes smaller than the copper roughness. Moreover, the thickness of the PCB is also a very important factor of consideration during material selection. For instance, at higher frequencies, thick materials are prone to significant radiation losses as compared to thin materials. But at the same time, thin materials are dominated by conductor losses. Therefore, choosing a thin material with lower copper surface roughness is highly recommended for use at mm-wave frequencies over a thick material with low conductor losses but higher radiation losses [121].

As applications move up in frequency, affordable and high-performance substrates are required. In military applications, polytetrafluoroethylene (PTFE) substrates have been considered one of the best materials for RF applications. Teflon or PTFE-based substrates provide very low $\tan\delta$ (0.0018 at 10 GHz) and offer great chemical resistant properties, small water absorption, and high-temperature resistance [123]. However, when the applications require high-performance substrates such as PTFE, those substrates present limitations in cost. For example, PTFE is ten times the price of epoxy glass (FR4) and five times the price of ceramic-based materials. In addition, PTFE requires extra preparation for good adhesion for the multi-layer PCB process. Moreover, conventional PTFE substrates are soft and the CTE of the conventional PTFE substrates is very high (180 to 205 ppm/°C)[123]. The proposed method enables the use of the existing inexpensive materials that provide excellent thermal properties and great compatibility with a multi-layer fabrication process to obtain high RF performance in terms of desirable ε_{eff} and $\tan\delta$ values for antenna applications.

4.3 Effective Permittivity and Mixing Techniques:

Analytical Models

Several dielectric mixing rules have widely been used in literature such as Maxwell Garnett mixing formula [124], Bruggeman [125], Looyenga [126], Polder-van Santen rule [127], and Lichtenecker mixing formula [128], etc. The most common of all these models is the Maxwell Garnett rule which is in the simplest form and is broadly applicable to a variety of particles. The theoretical background and analytic derivation of the Maxwell Garnett formula are not discussed here, as it is not the focus of this work. However, in this section, the required formulations that

can be used for finding the ε_{eff} of any isotropic (sphere, cubic) and anisotropic (cylindrical, disc, elliptical) particles, are presented.

The generic Maxwell Garnett mixing rule having the depolarization information is given by the expression: (4.1) [129]:

$$\varepsilon_{eff_j} = \varepsilon_2 \left(1 + \frac{(\varepsilon_1 - \varepsilon_2)f}{\varepsilon_2 + (\varepsilon_1 - \varepsilon_2)(1 - f)N_j} \right); j = x, y, z \quad (4.1)$$

where, ε_1 and ε_2 are the permittivities of the particle and the host material, respectively. f is the fractional volume of the mixing particle, and N_j is the depolarization factor opposing the direction of the applied field.

4.3.1 Isotropic Particles

For an isotropic particle, the value of $N_j = 1/3$, where $j = x, y$, and z [129]. An example geometry of the isotropic particle is shown in Figure 4.1(a). By replacing the value of $N = 1/3$ in (4.1), it can be easily shown that the ε_{eff} for an isotropic particle can be calculated from a more common form of Maxwell Garnett mixing formulation expressed as:

$$\varepsilon_{eff} = \varepsilon_2 \frac{\varepsilon_1 + 2\varepsilon_2 + 2f(\varepsilon_1 - \varepsilon_2)}{\varepsilon_1 + 2\varepsilon_2 - f(\varepsilon_1 - \varepsilon_2)} \quad (4.2)$$

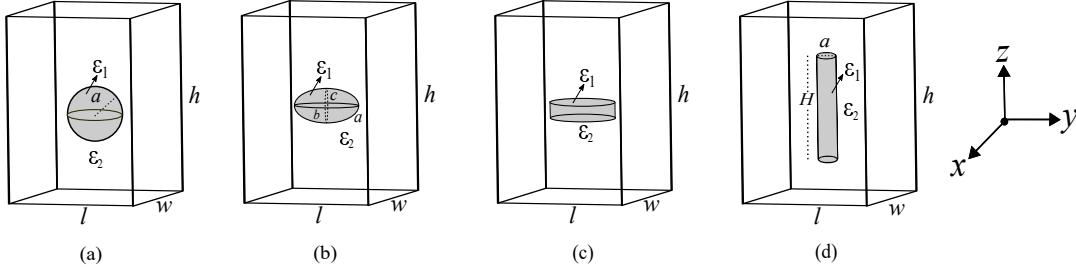


Figure 4.1: Isotropic and anisotropic unit cell geometries used for the mixing technique. (a) sphere (b) ellipsoid (c) disc (d) cylinder

4.3.2 Anisotropic Particles

- **Ellipsoid Particle**

The geometry of an ellipsoid inside a host medium is shown in Figure 4.1(b). The depolarization factor N_x can be calculated from the expression in (4.3)[130]:

$$N_x = \frac{abc}{2} \int_0^{+\infty} \frac{ds}{(s + a^2)\sqrt{(s + a^2)(s + b^2)(s + c^2)}} \quad (4.3)$$

where, a , b , and c are the semi-axes of the ellipsoid. N_y and N_z can be calculated by interchanging b and a , and c and a , respectively. Closed-form expressions of (4.3) can be found in [131]. The depolarization factors of a general ellipsoid can also be found in [132] -[133]. Once the depolarization factors are known, the equation of (4.1) can be used to find the ε_{eff} in any direction.

- **Disc Particle**

The unit cell geometry of a disc particle is shown in Figure 4.1(c). As the particle is symmetric in x - and y -directions, the permittivity in either x - or

y -, and z -directions need to be evaluated. Equations (4.4) and (4.5) can be used to find the ε_{eff} in the z - and x - or y -directions, respectively [134].

$$\frac{1}{\varepsilon_{effz}} = \frac{f}{\varepsilon_1} + \frac{1-f}{\varepsilon_2} \quad (4.4)$$

$$\varepsilon_{effxy} = f\varepsilon_1 + (1-f)\varepsilon_2 \quad (4.5)$$

• Cylindrical Particle

The geometry of the unit cell of the cylindrical particle is shown in Figure 4.1(d). The formulations for the cylindrical particle that is used in the design of our proposed material are discussed in detail. Unfortunately, the exact analytical models for finding the ε_{eff} of the material consisting of cylindrical particles do not exist. However, very few models discussing the approximate depolarization factor for the cylindrical particle exist, which are discussed in this section. In the following sections, the performance of these models based on their depolarization factors is compared. It is shown that after choosing the right depolarization factor, the generic Maxwell Garnett formulation is then applicable to find the ε_{eff} using the cylindrical particle, in any direction (x -, y -, z -) or any polarization ($\parallel$ or $\perp$) of the applied field with an overall error of $\leq 5\%$ in all cases.

The polarizability (α_j) for a cylindrical particle is calculated numerically using the method of moments (MoM) [135] -[136]. For the anisotropic particle, α_j for $j = z$ and $j = x$ or y are given by the expressions of (4.6) and

(4.7), respectively:

$$\alpha_z \approx 3.8662(\varepsilon_r - 1) \times \frac{\varepsilon_r^3 - 0.0519\varepsilon_r^2 + 0.9427\varepsilon_r + 1.5840}{\varepsilon_r^4 + 3.2226\varepsilon_r^3 - 0.0021\varepsilon_r^2 + 5.3391\varepsilon_r + 3.8662} \quad (4.6)$$

$$\alpha_{x,y} \approx 3.1691(\varepsilon_r - 1) \times \frac{\varepsilon_r^3 + 2.0283\varepsilon_r^2 + 1.9821\varepsilon_r + 1.5799}{\varepsilon_r^4 + 4.4453\varepsilon_r^3 + 6.2134\varepsilon_r^2 + 6.05\varepsilon_r + 3.1691} \quad (4.7)$$

where ε_r is the permittivity of the mixing particle. The depolarization factor N_j can then be found from α_j using (4.8) [129], which is then substituted in (4.1) to find the ε_{eff} .

$$\alpha_j = \left(\frac{\varepsilon_2(\varepsilon_1 - \varepsilon_2)f}{\varepsilon_2 + (\varepsilon_1 - \varepsilon_2)N_j} \right); j = x, y, z \quad (4.8)$$

The second formulation, known as Van Beek model [134], uses (4.1) for a prolate spheroid case with approximate N_j values and provides a final expression for calculating ε_{effz} is shown in (4.9):

$$\varepsilon_{effz} = \varepsilon_2 + \frac{f}{3} \frac{(\varepsilon_1 - \varepsilon_2)(5\varepsilon_2 + \varepsilon_1)}{\varepsilon_2 + \varepsilon_1} \quad (4.9)$$

Another famous formulation, known as Rayleigh's model for finding ε_{eff} in the direction perpendicular to the cylinder axes ($\varepsilon_{effx,y}$) is expressed in (4.10)[137]:

$$\varepsilon_{eff_{x,y}} = \varepsilon_2 \frac{\varepsilon_1 + \varepsilon_2 + f(\varepsilon_1 - \varepsilon_2)}{\varepsilon_1 + \varepsilon_2 - f(\varepsilon_1 - \varepsilon_2)} \quad (4.10)$$

The approximate value of $N_z = 1/2$ for a cylindrical particle in the z -direction is given by the Kittel model [138]. To obtain the most accurate results, this value is optimized slightly. It should be noted that all the mixing rules that exist so far, have certain limitations, either in terms of the f or the particle geometry. There is no single formulation that works best for any particle's shape or any f . However, some of these models provide a good starting point, and then after some optimization, close agreement between the simulated and analytical results can be obtained.

4.4 Effective Permittivity Analysis

This section shows the complete analysis for finding the ε_{eff} of both isotropic (spherical) and anisotropic (cylindrical) particles. Moreover, for each particle, two cases: conventional and complementary are considered. In a conventional case, the particle's permittivity ($\varepsilon_1 = 3.66$) is considered greater than the host permittivity ($\varepsilon_2 = 1$), i.e. $\varepsilon_1 > \varepsilon_2$. For the complementary case, the particle's permittivity ($\varepsilon_1 = 1$) is set lower than the host permittivity ($\varepsilon_2 = 3.66$), i.e. $\varepsilon_1 < \varepsilon_2$.

4.4.1 Isotropic Particle

Figures. 4.2(a) and (b) show the unit cell geometries for the conventional and complementary isotropic particles, respectively. The dimensions of the unit cell

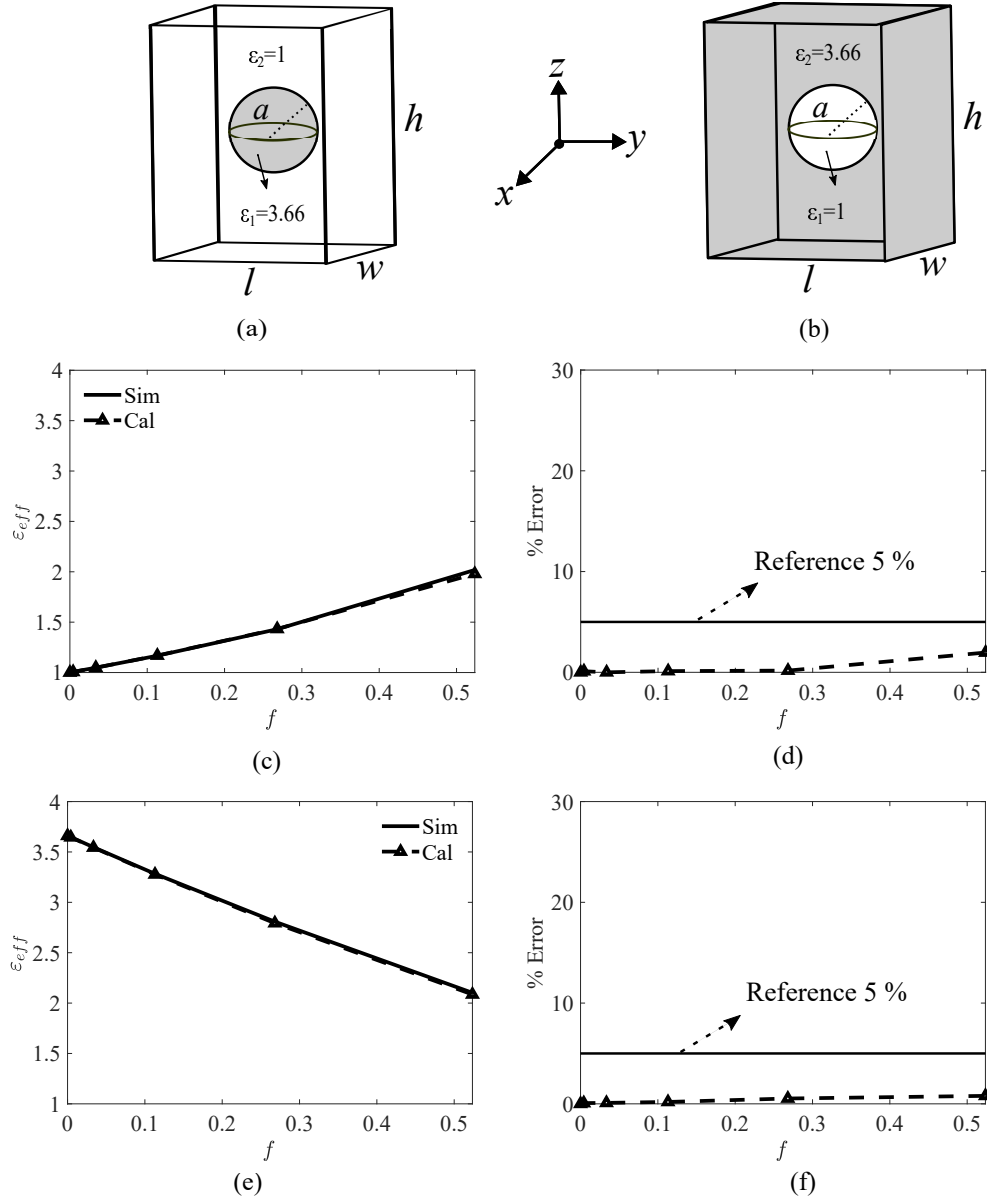


Figure 4.2: Permittivity analysis for the isotropic spherical particle. (a) Unit cell geometry of the conventional particle. (b) unit cell geometry of the complementary particle. (c) simulated and calculated ϵ_{eff} for the conventional particle as a function of the f . Calculated results use (4.2)[129]. (d) permittivity perceptual error (%) for the conventional particle as a function of f . (e) simulated and calculated ϵ_{eff} for the complementary particle as a function f . calculated results use (4.2)[129]. (f) permittivity perceptual error (%) for the complementary particle as a function of f .

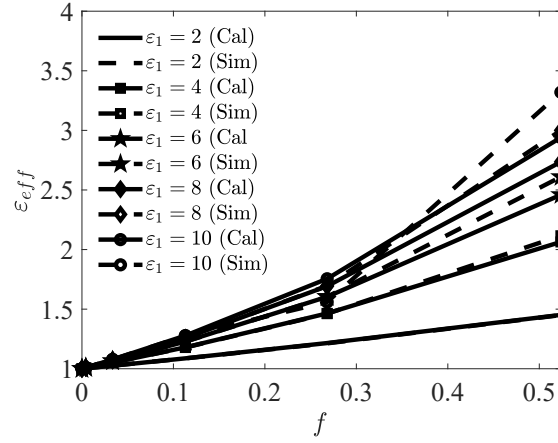
(length (l), width (w), and height (h)) are chosen as: $l = w = h = \lambda_g/10 = 4.5$ mm, at a frequency of 3.5 GHz, where λ_g is the guided wavelength given as: $\lambda_g = \lambda_o/\sqrt{\epsilon_r}$, and λ_o is the free-space wavelength. a is the radius of the sphere. As the particle has a full symmetry, the ϵ_{eff} remains the same irrespective of the direction of excitation as well as irrespective of the polarization of the incident field ($\parallel$ or $\perp$). The f of the spherical particle inside the host medium can be calculated as:

$$f = \frac{4\pi a^3}{3lwh} \quad (4.11)$$

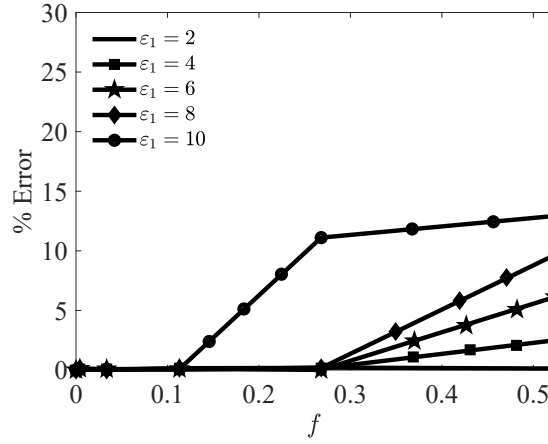
Different values of f are obtained by varying a . The range of f is given as: $f_{min} = 0$ (no particle inside), $f_{max} : a = (l = w = h)/2$ (touching condition).

Figures 4.2(c) and (e) show the simulated and calculated (using Eq. 4.2) ϵ_{eff} curves for the conventional and complementary cases, respectively. Figures 4.2(d) and (f) show the % error for both cases. It can be seen that the results are in very good agreement with an overall error of $< 2\%$ for the whole range of f . Also, it can be observed that the % error starts increasing for the large values of f . This is one of the widely discussed limitations of the Maxwell Garnet mixing rule (4.1) and (4.2).

The other limitation of these formulations is in terms of the difference (Δ) between ϵ_1 and ϵ_2 . This is shown in Figure 4.3, where it can be seen that by increasing the particle's permittivity ϵ_1 from 2 to 10, while $\epsilon_2 = 1$, the calculated results start diverging from the simulation ones. However, for the small values of f , a good agreement can be noticed. This analysis shows that the Maxwell Garnett mixing rule can still be used for large Δ only by considering very small values of f .



(a)



(b)

Figure 4.3: Limitation of Maxwell Garnett model (4.2)[129]. (a) Comparison of the simulated and calculated ε_{eff} for different values of ε_1 when $\varepsilon_2 = 1$. (b) permittivity perceptual error (%) as a function of f .

4.4.2 Anisotropic Particle

- ε_{eff} in the z -direction

- *Conventional Particle:*

To find the ε_{eff} in the z -direction (ε_{effz}), consider a cylindrical particle having $\varepsilon_1 = 3.66$ (RO4350B), embedded in the host medium of $\varepsilon_2 = 1$ as shown in Figure 4.4(a). The dimensions of the unit cell (l , w , and h)

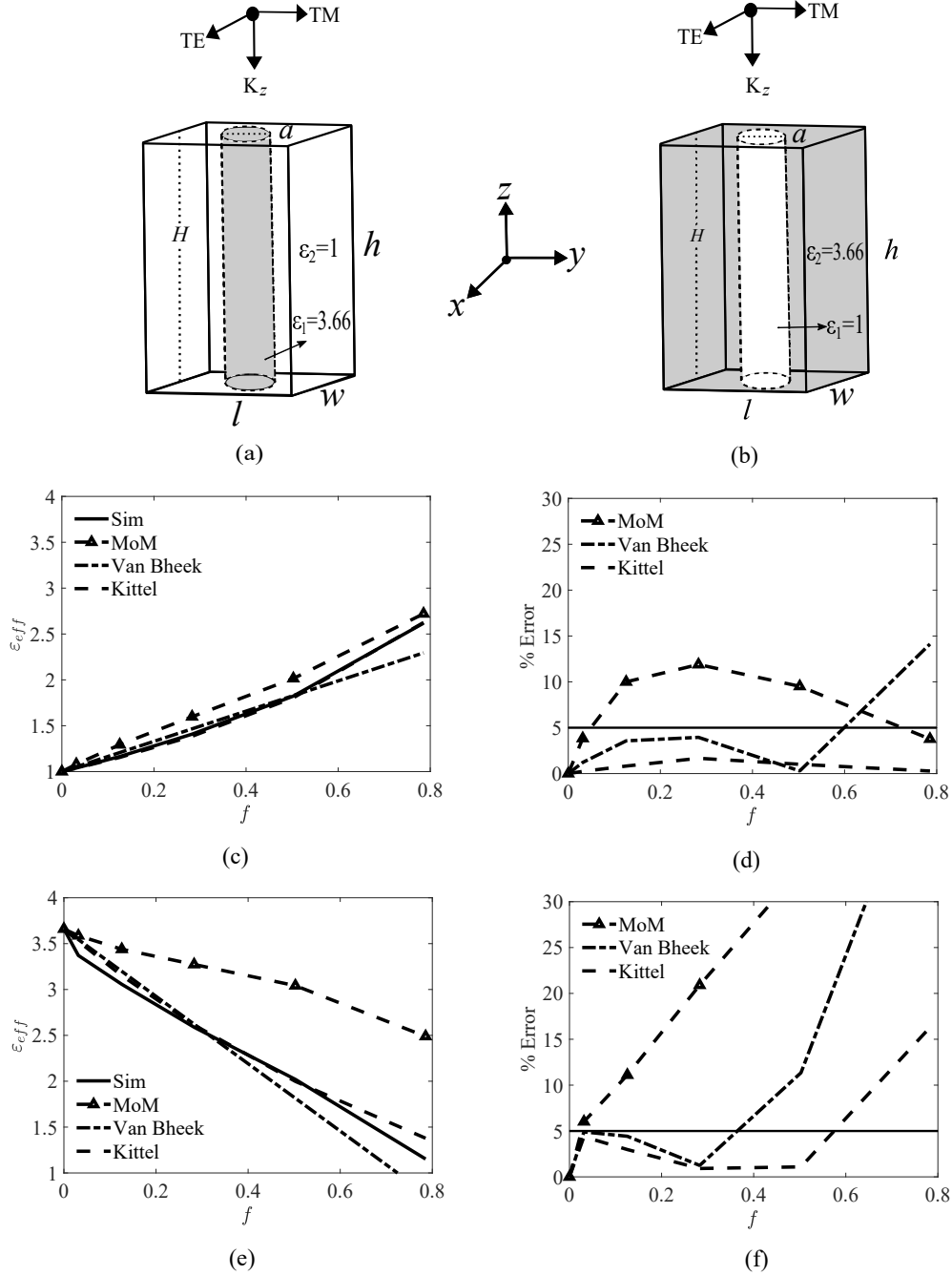


Figure 4.4: Permittivity analysis for the cylindrical particle excited in the z -direction. (a) unit cell geometry of the conventional particle. (b) unit cell geometry of the complementary particle. (c) comparison of the models: MoM (4.6)[135], Van Bheek (4.9)[134], Kittel ($N_z = 1/2$)[138], with the simulation results, for the conventional particle. (d) permittivity perceptual error (%) in the models for the conventional particle. (e) comparison of the models: MoM (4.6)[135], Van Bheek (4.9)[134], Kittel ($N_z = 1/1.85$), with the simulation results, for the complementary particle. (f) permittivity perceptual error (%) in the models for the complementary particle.

are chosen as: $l = w = \lambda_g/10$, and $h = 2.97$ mm, at a frequency of 3.5 GHz. H is the height of the particle, $H=h$. The particle is excited in the z -direction using Floquet ports with Master & Slave boundaries in HFSSTM. As the direction of propagation (K_z) is parallel to the axis of the cylinder, the polarization of the incident E-field ($\parallel$ or $\perp$) does not affect the ε_{eff} due to the symmetry of the structure. The f of the particle is given as:

$$f = \frac{\pi a^2 H}{4lwh} = \frac{\pi a^2}{4lw} \quad (4.12)$$

where a is the diameter of the cylinder. Figure 4.4(c) shows the comparison of the models with the simulation results for ε_{effz} , considering a conventional particle, and Figure 4.4(d) shows the % error. It can be seen that the value: $N_z = 1/2$ from the Kittel model is in good agreement with the simulation results with an error of $< 5\%$ for the whole range of f .

– ***Complementary Particle:***

Figure 4.4(b) shows the unit cell geometry of the complementary cylinder particle having $\varepsilon_1=1$ inside a host medium having $\varepsilon_2=3.66$. The complementary particle having $\varepsilon_1=1$ is represented by a perforation (a through-drill where $H=h$). Figures 4.4(e) and 4.4(f) show the ε_{eff} and % error, respectively, for the complementary particle. It can be seen that for a very small value of $f < 0.03$, the models show good agreement. However, for the large values of f , the MoM model presents a significant departure as compared to the other two models. This is

because the α_j values in the MoM model were calculated for $\varepsilon_1 > \varepsilon_2$. The N_z value given in the Kittel model is slightly tuned to $N_z = 1/1.85$ to get an error of $< 5\%$ for around 58% of f .

- **ε_{eff} in the x -, y -directions**

To evaluate the ε_{eff} in x - or y -direction, it is important to note that the particle is symmetric in both x - and y -directions, therefore, $\varepsilon_{effx} = \varepsilon_{effy}$. However, the polarization of the incident field: $\mathbf{E}_{\parallel}$ or $\mathbf{E}_{\perp}$ as shown in Figure 4.5, has an impact on the polarization field created inside the particle, which ultimately affects the ε_{eff} . Hence, the $\varepsilon_{eff\parallel}$ and $\varepsilon_{eff\perp}$ are evaluated separately, for both conventional and complementary cases.

- ***Conventional Particle:***

Figure 4.5 shows the geometry of the conventional particle as well as the comparison of the models for both $\parallel$ and $\perp$ polarizations. It can be seen that the MoM model presents a minimum of 10% error for both cases. The Rayleigh model shows a comparatively large deviation in the $\parallel$ case while its error is $< 5\%$ for the whole f , in the $\perp$ case. The initial value of $N_z = 1/2$ provided by the Kittel model is tuned to have an error of $< 5\%$ for both cases. This analysis shows the sensitivity of ε_{eff} to the polarization of the incident field.

- ***Complementary Particle:***

Figure 4.6 shows the geometry of the complementary particle and the ε_{eff} analysis for both $\parallel$ and $\perp$ polarizations. It can be seen that, as expected the MoM models show significant deviation from the simulation results for both $\parallel$ and $\perp$ cases. The Rayleigh model shows a maximum of 12% error for the $\parallel$ case while it converges well for the $\perp$ case with

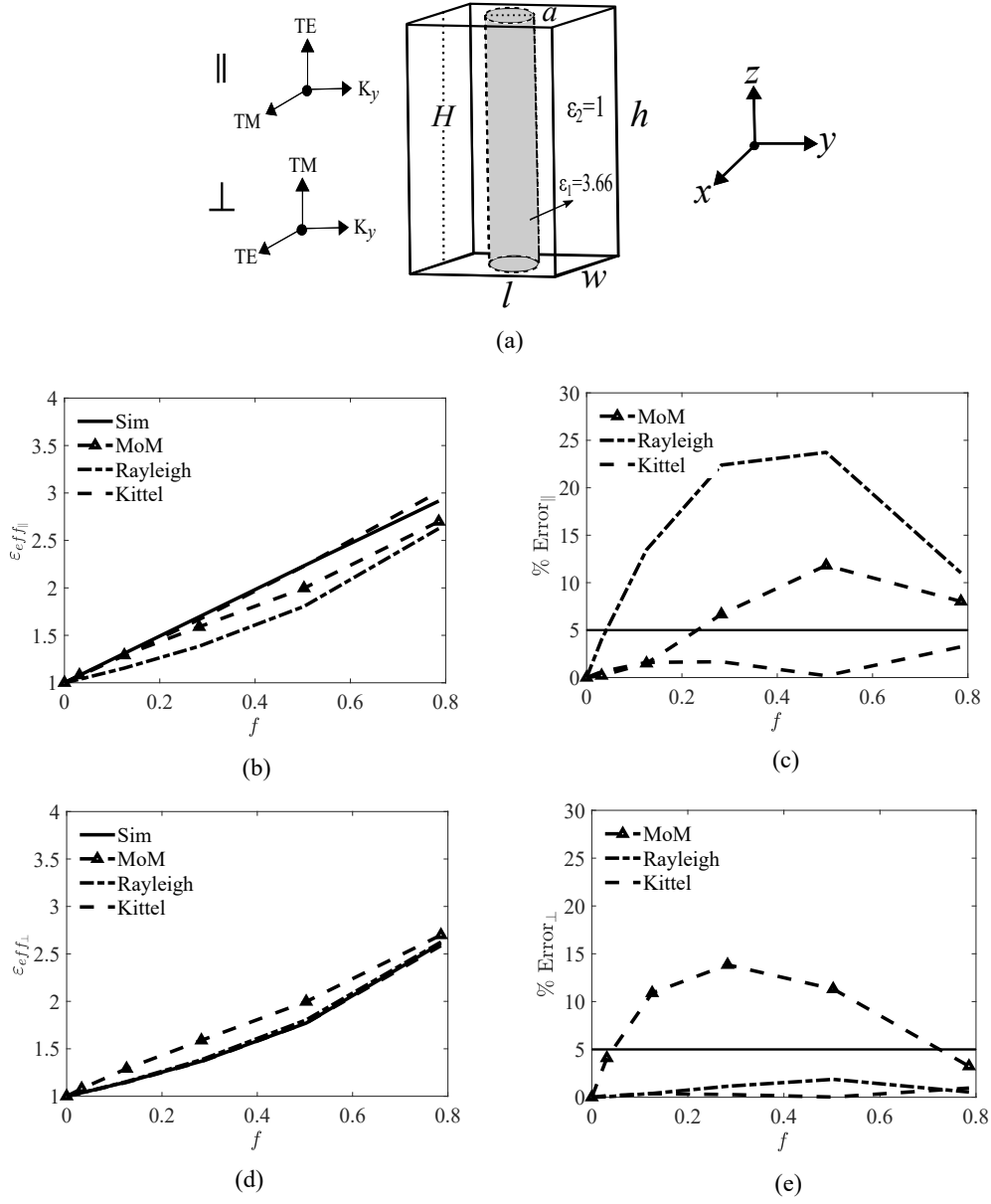


Figure 4.5: Permittivity analysis for the conventional particle excited in the x -, y -directions. (a) unit cell geometry of the conventional particle. (b) comparison of the models: MoM (4.7)[136], Rayleigh (4.10)[137], Kittel ($N_z = 1/15$), for $\epsilon_{eff||}$. (c) permittivity perceptual error (%) error in the models for $\epsilon_{eff||}$. (d) comparison of the models: MoM (4.7)[136], Rayleigh (4.10)[137], Kittel ($N_z = 1/1.8$), for $\epsilon_{eff\perp}$. (e) permittivity perceptual error (%) in the models for $\epsilon_{eff\perp}$.

an error of $< 5\%$ for the whole f . The values of N_z in the Kittel model are tuned for both cases to have an error of $< 5\%$.

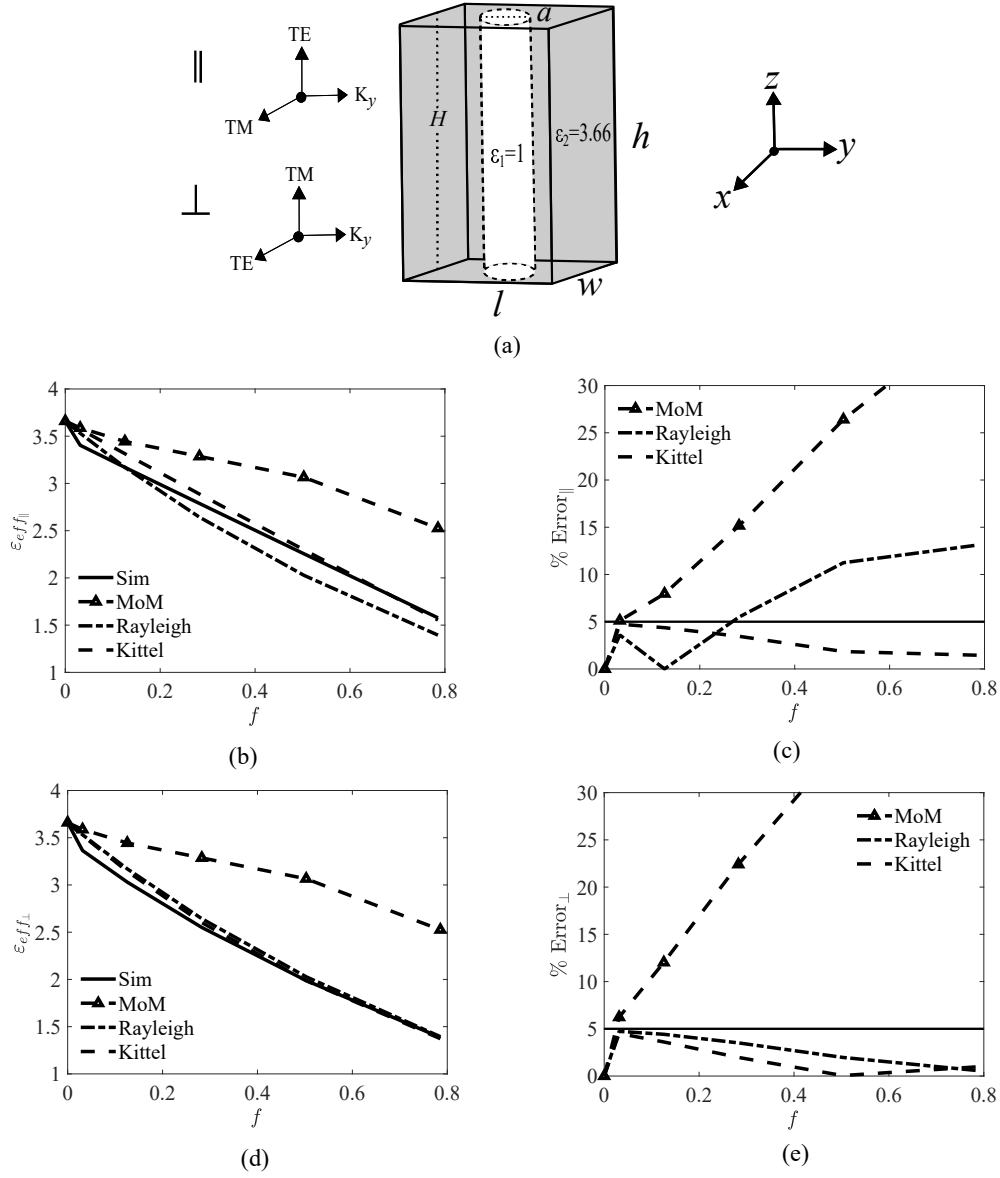


Figure 4.6: Permittivity analysis for the complementary particle excited in the x -, y -directions. (a) unit cell geometry of the complementary particle. (b) comparison of the models: MoM (4.7)[136], Rayleigh (4.10)[137], Kittel ($N_z = 1/20$), for $\epsilon_{eff||}$. (c) permittivity perceptual error (%) error in the models for $\epsilon_{eff||}$. (d) comparison of the models: MoM (4.7)[136], Rayleigh (4.10)[137], Kittel ($N_z = 1/1.8$), for $\epsilon_{eff\perp}$. (e) permittivity perceptual error (%) in the models for $\epsilon_{eff\perp}$.

4.5 Material Design and Synthesis: Microwave and mm-Wave Frequencies

4.5.1 Material Requirements

Among the several critical parameters discussed in the introductory section for choosing the right material for both microwave and mm-Wave applications, the ε_r and the material's dielectric loss or $\tan\delta$ are of significant importance. As compared to the microwave frequencies, where a high ε_r can be used, at mm-Wave frequencies, a low ε_r material is highly recommended because using a high ε_r material would further reduce the size of the structure thus creating more challenges in the fabrication [139]. Also, the losses in the material are often related to the value of ε_r . The higher the value of ε_r , the higher the losses in the material [139]. Therefore, at mm-Wave frequencies where the losses are significant, a low ε_r material is used to reduce further losses in the material. Moreover, at high frequencies, the speed of signal propagation in the high-speed circuits is very important. The signal propagation delay (t_d) depends on both the guiding structure as well as on the ε_r of the material and is expressed as (4.13)[122]:

$$t_d = \sqrt{\varepsilon_r \times L \times C} \quad (4.13)$$

where L and C are the inductance and capacitance of the line. Eq. (4.13) shows that by using a low ε_r material or different transmission line structures (microstrip, stripline or coplanar waveguide, etc.), t_d can be directly reduced. Cross-talk is also of utmost consideration when referring to mm-Wave frequencies. Using a low ε_r material in the high-frequency circuits can decrease the capacitive coupling between the conductors and thus cross-talk can be reduced [122]. Aside from the

values of ε_r , more important is the consistency of the ε_r at mm-Wave frequencies. A slight variation in material ε_r can cause a change in the impedance of the transmission line which can cause unexpected variations in the phase angles resulting in errors in radar signal detection [121].

Another important property of the material is the $\tan\delta$. At both microwave and mm-Wave frequencies, dielectrics with low losses are used to reduce attenuation and heating in the circuit [122]. Current substrate materials have $\tan\delta$ values between $0.002 < \tan\delta < 0.02$. However, At the mm-Wave frequencies, the requirements become more stringent, and $\tan\delta$ values of < 0.002 are desired [139].

4.5.2 Design Trade-offs

As compared to other particles previously discussed in Sec. 4.4, the complementary cylindrical particles inside a host medium which are commonly known as air perforations, offer several advantages in terms of easiness in fabrication (can be easily drilled) and measurement, reduce complexity in modeling the depolarization field due to its perfect alignment inside the structure, and help in the reduction of the $\tan(\delta)$ of the material. The air perforations in the host material with $\varepsilon_1 \approx 1$ and a $\tan\delta \approx 0$, have a very useful feature of reducing the overall $\tan\delta$. Figure 4.7 shows the reduction in the $\tan(\delta)$ of the host material by using a complementary cylindrical particle. It can be seen that when there is no particle inside i.e. $f = 0$, the original value of $\tan\delta = 0.004$ (RO4350B). However, for the maximum value of $f = 0.78$, the value of $\tan\delta = 0.0012$, which shows a reduction of around 70% in the $\tan\delta$ of the core material.

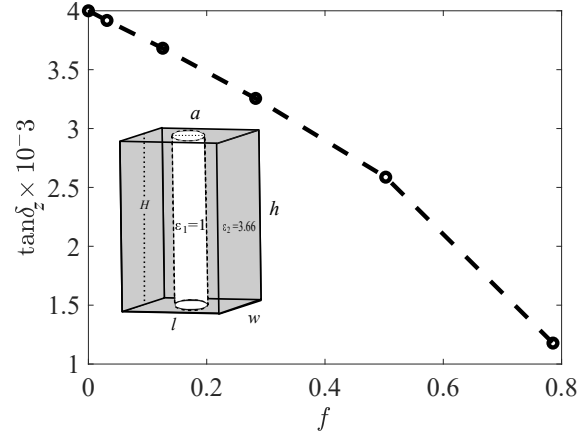


Figure 4.7: Reduction in the overall $\tan(\delta)$ of the material based on different values of f .

4.5.3 Design Procedure

Figure. 4.8 shows the simulation setup for the unit cell geometry of the proposed material structure using an infinite array approach in HFSSTM. The required Floquet ports along with the Master and Slave boundaries are used. In this case, the material is excited in the z -direction with two de-embedded ports. The real and imaginary parts of the S-parameters are used to extract the constitutive parameters (ϵ_{eff} , μ_r , $\tan\delta$) by using the Smith algorithm [140].

4.5.4 Size of the Unit Cell

Determining the size of the unit cell of the periodic structure in terms of the wavelength is critical for extracting the correct values of the constitutive parameters [141]. For this, the particle should meet the effective-homogeneity condition which is when the size of the unit cell is $< \lambda_g/4$ [141]. Moreover, a slight change in the value of ϵ_r over the range of frequencies can degrade the performance specifically at mm-Wave frequencies. Figure 4.9 shows the effect of the size of the unit

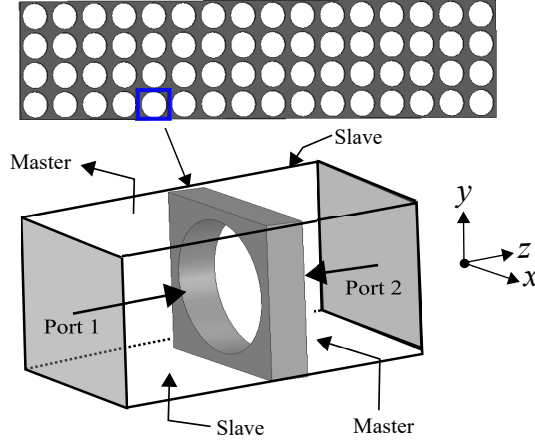


Figure 4.8: Simulation setup showing the unit cell geometry of the periodic structure with Floquet ports excitation and boundaries.

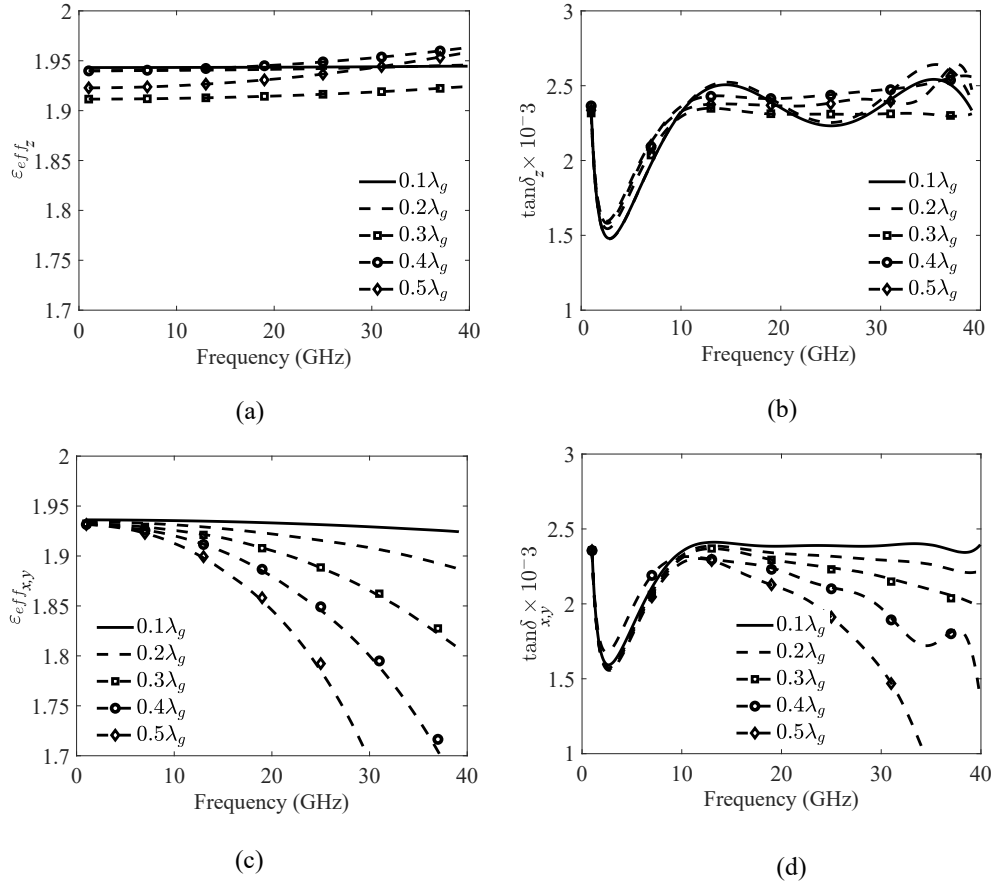


Figure 4.9: ε_{eff} and $\tan\delta$ as a function of the unit cell dimensions for the complementary cylindrical particle using 4350B with perforation. (a) ε_{eff} in the z -direction. (b) $\tan\delta$ in the z -direction. (c) ε_{eff} in the x -, y -directions. (d) $\tan\delta$ in the x -, y -directions.

cell on the ε_{eff} and $\tan\delta$ over the range of frequencies, for z - and x -, y -directions. For the purpose of illustration, a cubic unit cell of a complementary cylindrical particle excited in the z - and x - or y -directions is considered with $f = 56\%$. Different dimensions of the unit cell are considered in terms of the λ_g at the highest frequency of operation i.e. 40 GHz. The cases considered have the dimensions $\leq 0.5\lambda_g$, because, for the size of the unit cell greater than $0.5\lambda_g$, no convergence was observed. For comparison, f should remain the same for all the cases. Therefore, for each case, the diameter of the cylinder is adjusted to get the same values of f .

Figures. 4.9(a) and (b) show the ε_{eff} and the $\tan\delta$ in the z -direction. Figure 4.9(a) shows the difference in the values of ε_{effz} at lower frequencies as well as an increasing trend at higher frequencies for different dimensions of the unit cell. However, for the case of $0.1\lambda_g$, the values of ε_{effz} are stable over the entire frequency range. Figures 4.9(c) and (d) show this analysis for the x -, y -directions. It can be seen that for the unit cell size of $0.1\lambda_g$, stable performance is obtained. However, when the size gets larger, a decreasing trend can be observed. From this, it can be concluded that in order to have a steady performance in terms of ε_{effz} or $\varepsilon_{effx,y}$, the z - or x -, y -dimensions of the unit cell should be at least $\leq 0.1\lambda_g$ at the higher frequency of operation. The smaller the size of the unit cell, the more stable the performance. However, fabrication tolerances should be considered while determining the size of the unit cell.

The results from this analysis are summarized in Table 4.1. The dimensions of the unit cell are $0.1\lambda_g$ at 40 GHz. As the reference RO4350B sample is isotropic, the values of ε_{eff} and $\tan\delta$ for any direction are the same given in the datasheet. However, for the proposed sample loaded with anisotropic complementary cylindrical particle, the values of ε_{eff} and $\tan\delta$ slightly differ between the x -, y -, and z -directions. The reason for this has been explained in detail in Sec. 4.4.

Table 4.1: Summary of ε_{eff} and $\tan\delta$ for x -, y -, and z -directions for solid and proposed material for 1 GHz to 40 GHz.

	Freq.	Size	Results								
	(GHz)	(λ_g^3)	Parameters				Reduction factor				
			f	$\varepsilon_{effx,y}$	ε_{effz}	$\tan\delta_{x,y}$	$\tan\delta_z$	$\varepsilon_{effx,y}$	ε_{effz}	$\tan\delta_{x,y}$	$\tan\delta_z$
Solid Reference 4350B	40	$0.1\lambda_g$	–	3.66	3.66	0.004	0.004	–	–	–	–
Proposed	40	$0.1\lambda_g$	0.56	1.92	1.95	0.0022	0.0024	47.5%	46.7%	45%	40%

For determining the size of the unit cell for the W-band, the same unit cell size analysis was carried out at 77 GHz, using solid RO3003 as the reference sample. The dimensions of the unit cell were chosen based on the stable performance as well as keeping the fabrication tolerances in consideration. As the permittivity is measured in the z -direction, for sample-3 the z -dimension or thickness H of the material, which is more sensitive to the performance, was chosen to be small as $0.226\lambda_g$ at 77 GHz in order to achieve accurate and stable results. The other dimensions of the unit cell were kept comparatively larger as $1\lambda_g$ due to fabrication restrictions.

4.5.5 Results and Discussion: ADM for Antennas

To validate the analysis presented before, two different samples were fabricated and measured in the S-band at 3.5 GHz and a third sample in the mm-Wave W-band at 77 GHz. The proposed samples were designed using the analysis presented in the previous sections. These samples were fabricated using the LPKF Protomat S103 machine and were measured using the setups shown in Figure 4.10. The ε_{eff} of the S-band samples: sample-1 and sample-2 were measured using the waveguide setup of Figure 4.10 (a), and the W-band sample-3 was characterized using the mm-Wave RF scanner shown in Figure 4.10 (c).

The measurement results were compared with the reference samples. The

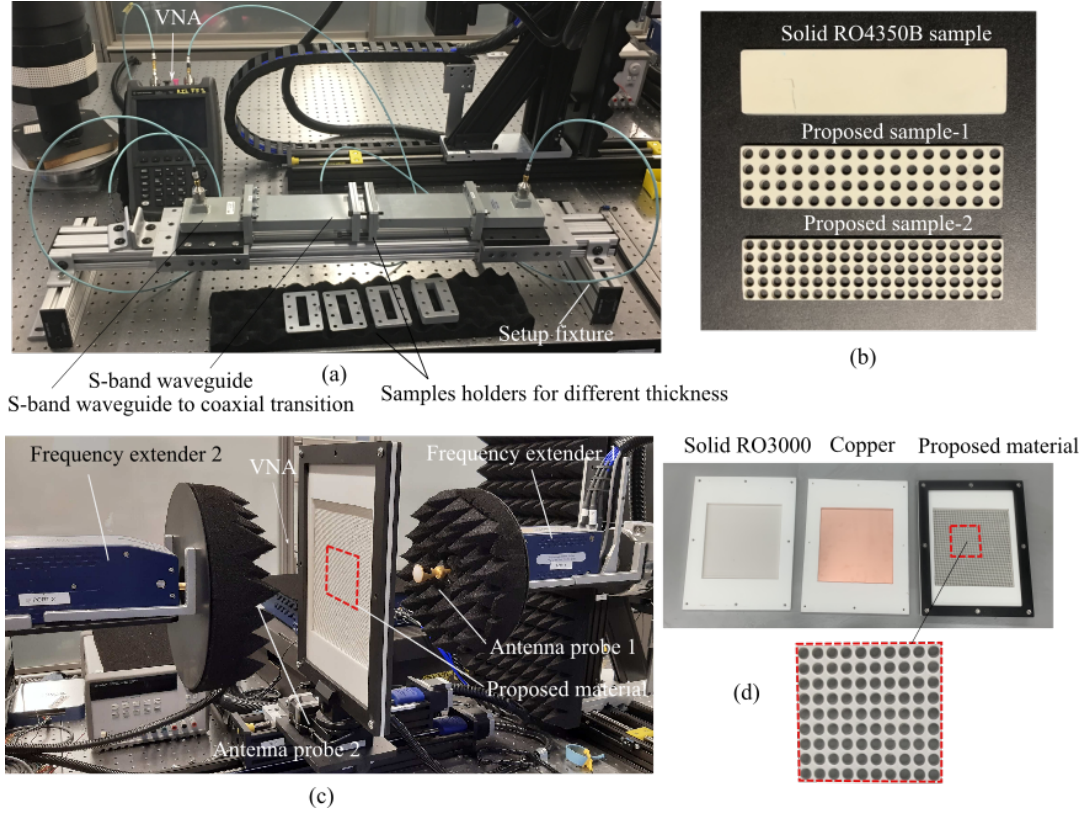


Figure 4.10: Material measurement setups for microwave and mm-Wave frequencies. (a) photo of the measurement setup for the S-band, based on the waveguide method. (b) material samples used for the S-band test. (c) photo of the measurement setup for the W-band, based on the free-space Gaussian beam method. (d) material samples used for W-band test.

simulated, calculated, and measured results in terms of ϵ_{effz} and $\tan\delta$ are shown in Figure 4.11 for the S-band samples, and in Figure 4.12 for the W-band sample-3. Figure 4.10(b) and (d) show the reference and the proposed samples for the S-band and the W-band, respectively. For the S-band samples: sample-1 and sample-2, the dimensions of the unit cell are kept $\leq \lambda_g/10$ at 3.5 GHz based on the analysis presented in Sec. 4.4. The dimensions of the unit cell in both cases are varied to obtain different values of f . Figures 4.11 and 4.12 show an excellent agreement between the simulated, calculated, and measured results. For sample-3, the thickness H of the material is chosen to be as small as $0.226\lambda_g$ at 77 GHz in

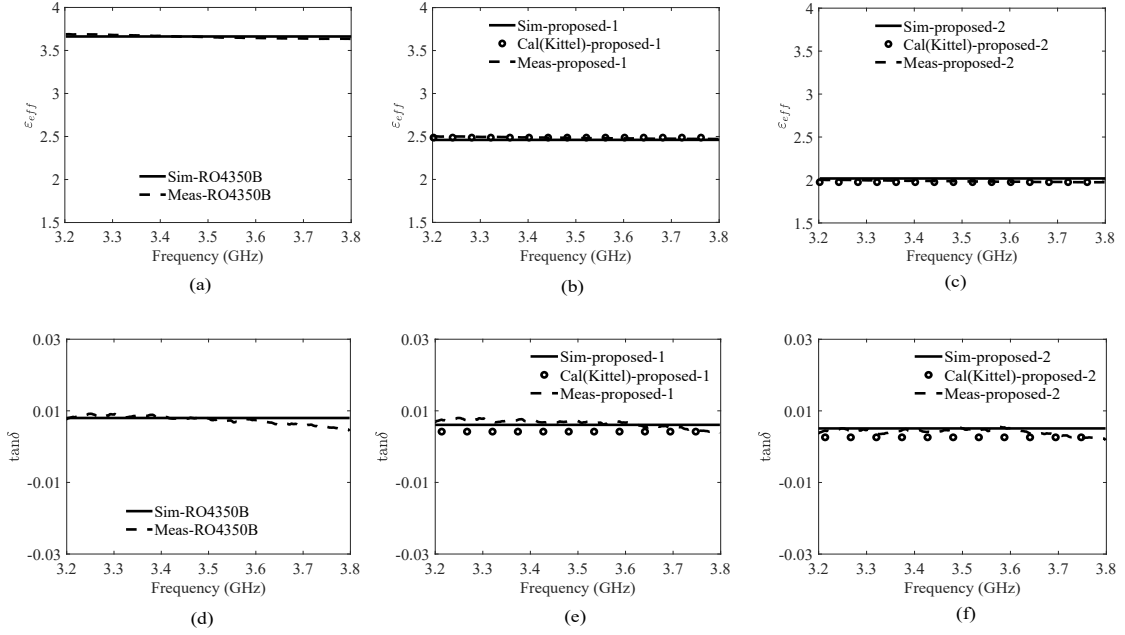


Figure 4.11: Comparison between measured, simulated, and calculated results in the S-band. (a) ε_{eff} solid sample (RO4350B). (b) ε_{eff} proposed sample-1. (c) ε_{eff} proposed sample-2 with Kittel ($N_z = 1/2.4$). (d) $\tan\delta$ solid RO4350B sample. (e) $\tan\delta$ proposed sample-1 with Kittel ($N_z = 1/2.8$). (f) $\tan\delta$ proposed sample-2 with Kittel ($N_z = 1/2.4$).

order to achieve accurate and stable results. The other dimensions of the unit cell are kept comparatively large due to fabrication challenges at 77 GHz. However, all the dimensions are tuned to get convergence in the results.

All the dimensions as well as the results for the S-band and the W-band samples are also summarized in Table 4.2. It should be noted that the measured value of the $\tan\delta$ for the reference RO4350B sample given in Table 4.2 is 0.008 and not 0.004 (from the datasheet). Similarly, the values of ε_r and $\tan\delta$ for the reference RO3003 sample are 2.91 and 0.0072, respectively, measured at 77 GHz. This variation in the results can be attributed to the selection of the measurement method, frequency of operation, external environmental conditions, and slight inaccuracy in the measurement setup. It should also be noted that each measured ε_{eff} curve

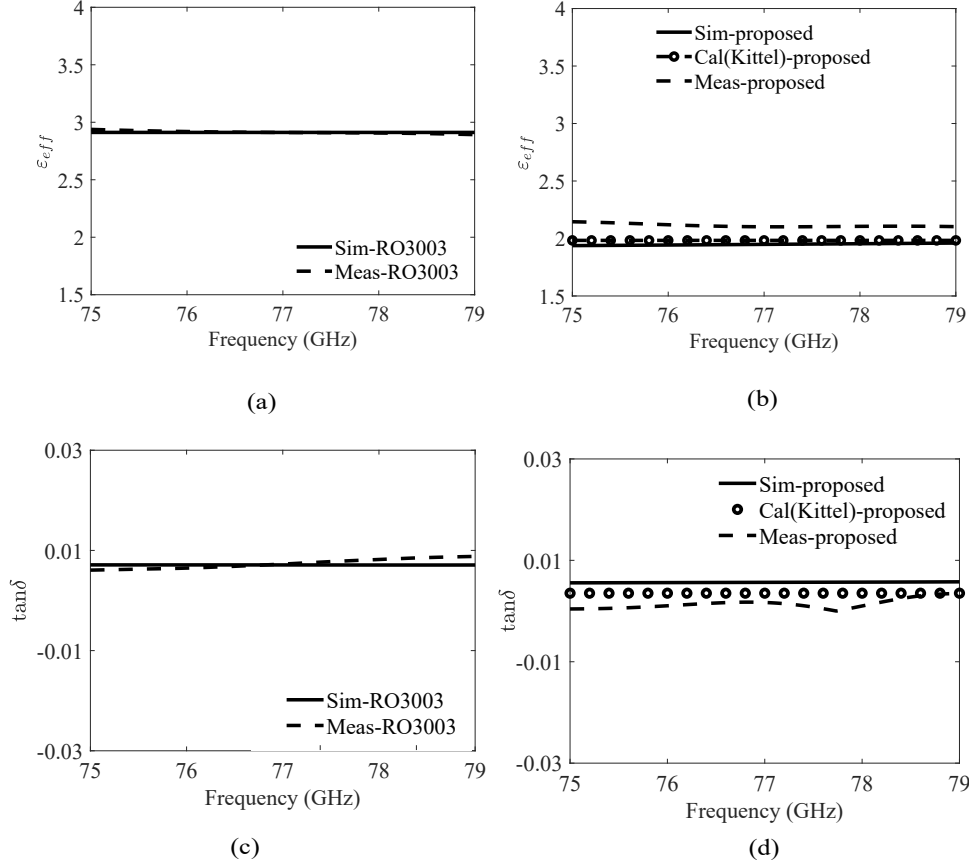


Figure 4.12: Comparison between measured, simulated, and calculated (Kittel model $N_z = 1/30$) results in W-band. (a) ϵ_{eff} solid sample (RO3003). (b) ϵ_{eff} proposed sample-3. (c) $\tan\delta$ solid sample (RO3003). (d) $\tan\delta$ proposed sample-3.

in Figure 4.11 or 4.12 represents the statistical average of several repeated measurements to ensure stability in the results and to reduce errors associated with the measurement. It can be seen that for sample-1 having $f = 37\%$, the reduction in ϵ_{effz} and $\tan\delta$ is 33% and 25%, respectively. For sample-2 with $f = 55\%$, a reduction of 45% and 38% is obtained in the ϵ_{effz} and $\tan\delta$, respectively, as compared to the solid reference sample RO4350B shown in Figure 4.11(a). Similarly, for sample-3 with $f = 50\%$, a reduction of 32% and 72% is achieved in the values of ϵ_{eff} and $\tan\delta$, respectively, as compared to the reference RO3003 sample shown in Figure 4.12(a).

Table 4.2: Summary of the measured results of ε_{eff} and $\tan\delta$ in the z -direction for solid and proposed material in the S-band and W-band.

	Material	Freq. (GHz)	Size					Results				
			(mm)					Parameters			Reduction factor	
			l	w	h	H	a	f	ε_{effz}	$\tan\delta_z$	ε_{effz}	$\tan\delta_z$
Reference	RO4350B	3.5	—	—	—	2.97	—	—	3.66	0.008	—	—
Sample-1	RO4350B	3.5	4.51	4.25	2.97	2.97	3	0.37	2.46	0.006	33%	25%
Sample-2	RO4350B	3.5	3.8	3.40	2.97	2.97	3	0.55	2	0.005	45%	38%
Reference	RO3003	77	—	—	—	0.508	—	—	2.91	0.0072	—	—
Sample-3	RO3003	77	2.25	2.25	0.508	0.508	1.8	0.5	1.98	0.002	32%	72%

Table 4.3: Performance comparison of the proposed work with other related works.

References	Theory & modeling	Design procedure	Particle geometry	Polarization & excitation analysis	Error analysis	Application	Freq. (GHz)
[114]	*	*	Cylindrical	*	*	SIIG	94
[115]	*	*	Cylindrical	*	*	SIIG	94
[116]	*	*	Cylindrical	*	*	SIIG	94
[117]	Yes	Yes	Cylindrical	*	*	Reconfigurability	4.2-4.9
[118]	*	*	Cylindrical	*	*	mm-Wave imaging	94
[119]	*	*	Cylindrical	*	*	mm-Wave	100
This work	Yes	Yes	Any isotropic or anisotropic	Yes	Yes	Any	1-110

* Not discussed in the paper.

Table 4.3 shows the performance comparison of the proposed work with other related works. It can be noticed that as compared to other works, the proposed work validates the analytical models based on the error analysis and also discusses the anisotropy of the material based on the direction of excitation and polarization analysis. Moreover, the proposed design procedures can be applied in the design of a high-performance material at any frequency between 1-110 GHz, for many applications.

4.6 Material Design and Synthesis: THz frequencies

Characterization of materials is a hot topic in the current research community. The constitutive parameters ε_r and $\tan\delta$ are frequency dependent, and their performance is not constant at higher frequencies [19]. Thus, proper characterization

techniques are required to analyze the behavior and losses in the materials at such high frequencies. Also, for antennas, low-loss materials with small values of ε_r between 2 – 4 are desired at such higher frequencies [19]. The lower the value of ε_r , the lower the capability of a material to store energy. Moreover, at THz frequencies, $\tan(\delta)$ values of <0.002 are usually used [139]. The thickness of the substrate material is also another important parameter in choosing the materials for high frequencies. Thick substrates usually suffer from surface wave losses, and thus, it is recommended to choose thin substrates at high frequencies.

This section shows the applicability and potential of the proposed particle mixing technique at THz frequencies. It is well-known that fabricating high-frequency antennas on thin substrates using standard PCB fabrication is not a feasible option. It imposes constraints on achieving finer size resolutions, potentially resulting in a lower limit for trace/space dimensions. This limitation in PCB fabrication comes from the fact the chemical wet-etching technique cannot properly etch a thick copper having a thickness of $17.5\ \mu m$ (1/2 oz copper), and $9\ \mu m$ (1/4 oz copper). Currently, copper cladding lower than $9\ \mu m$ is not commercially available. To overcome this, a solution that is usually discussed in literature, for instance, presented in [142], where at 165 GHz, traditional PCB etching is used when a $5\ \mu m$ thick copper is deposited on a $50\ \mu m$ thick LCP organic substrate. The fabrication issue can be solved by using micro-fabrication techniques. Silicon (Si) or silicon-germanium (SiGe) based semiconductor micro-fabrication techniques are commonly employed where the fabrication limitation can go to a few nanometers.

Another reason to go for Si semiconductor fabrication is the growing demand for fully integrated systems in which the antenna is integrated with other mm-Wave RF components. There are two widely and commercially used integration techniques: antenna-on-chip (AoC) and antenna-in-package (AiP). In AoC, the

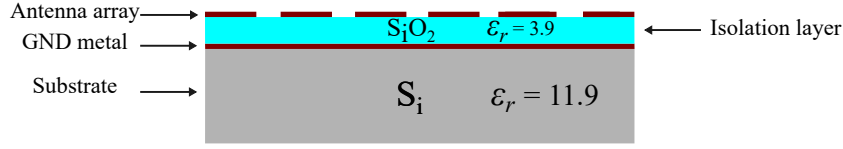


Figure 4.13: Cross-sectional view of a GND-shield technique in AoC technology on a Si substrate.

antenna is integrated with other front-end components on the same chip, while in AiP, the antenna is realized on a different substrate such as LCP or 3D glass, and then integrated with Si-based components.

In AoC, since the antenna is integrated on the same chip using a Si-wafer, this eliminates the component's interconnection losses and also reduces the overall size and cost of the system. However, Si which has a high ϵ_r , antennas implemented on a Si substrate have very low radiation efficiency as well as low gain and directivity.

4.6.1 Solutions Available in the Literature

- Several solutions exist in the literature to improve the gain and efficiency of such integrated antennas. One of the techniques usually used in literature is to isolate the lossy Si substrate by using a metallic ground (GND) shield as shown in Figure 4.13. Using this, the energy or radiation is reflected back and not absorbed in the Si substrate. In such a case, the antenna is then printed on a SiO₂ substrate having $\epsilon_r = 3.9$. The issue with this technique is that using the GND plane limits the thickness of the SiO₂ substrate and in most cases, its thickness is between 9-15 μm . Using such a thin SiO₂ substrate for the antenna (around 1% of the guided wavelength) creates

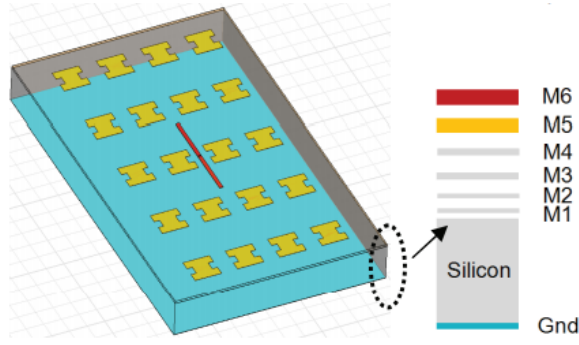


Figure 4.14: Dipole antenna over a HIS to improve gain [143].

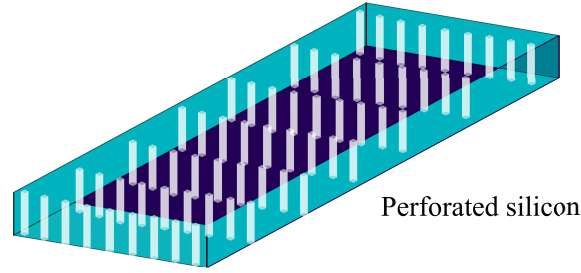


Figure 4.15: Illustration of the silicon substrate when loaded with air-cylindrical particles (perforated silicon).

ground image currents which limits the bandwidth and in such a case the ohmic losses also get much stronger which affects the radiation efficiency and gain of the AoC. [143].

- Another solution that is discussed in the literature is to use a high impedance surface (HIS) below the antenna (HIS is on layer M5 while dipole is on layer M1) as shown in Figure 4.14. The HIS is mainly used to improve the radiation efficiency and gain of the antenna. However, with this technique, the gain achieved is still low around 1 dBi as reported in [143]. Moreover, HIS is a narrow-band technique and does not show improvements in performance over a wide bandwidth [142].

4.6.2 Proposed Solution: Using ADM

By using the proposed ADM, ε_r and $\tan\delta$ of the portion of the silicon substrate on which the antenna or antenna array is to be realized, can be lowered to improve the efficiency and gain of the AoC. This section shows that the particle mixing technique can be extended to the sub-THz region for both antennas and filters on the same Si substrate. A Si substrate loaded with the perforated air-cylindrical particle is shown in Figure 4.15. Figure 4.16 shows a conceptual diagram of a future mobile phone having in-package or on-chip modules for low frequencies 3G/4G communication as well as semiconductor-based high frequencies on-chip modules for 5G/6G wireless communication. Figure 4.16(b) shows the detailed view of the 5G/6G module where antenna arrays and other RF components such as a filter can be integrated on the same silicon substrate. The ε_r for the antenna part can be tuned to as low as 2 from the initial value of 11.9, while for the filter the ε_r can be reduced to a value of 5.

4.6.3 Results and Discussion: ADM for Si-based Fabrication

Figure 4.17 shows the unit cell simulation of the proposed ADM in the Z-direction. The excitation is done through Floquet ports with Master and Slave boundaries in HFSSTM. The length (l) and width (w) of the unit cell are kept $\lambda_g/10 = 87 \mu m$ at 100 GHz while the height (h) is kept the same as the thickness of the silicon layer which is $200 \mu m$. The diameter (a) of the particle is then adjusted to achieve the desired f (ratio of the volume of the cylindrical particle over the total volume of the unit cell). The fractional volume (f_v) for the cylindrical particle can be

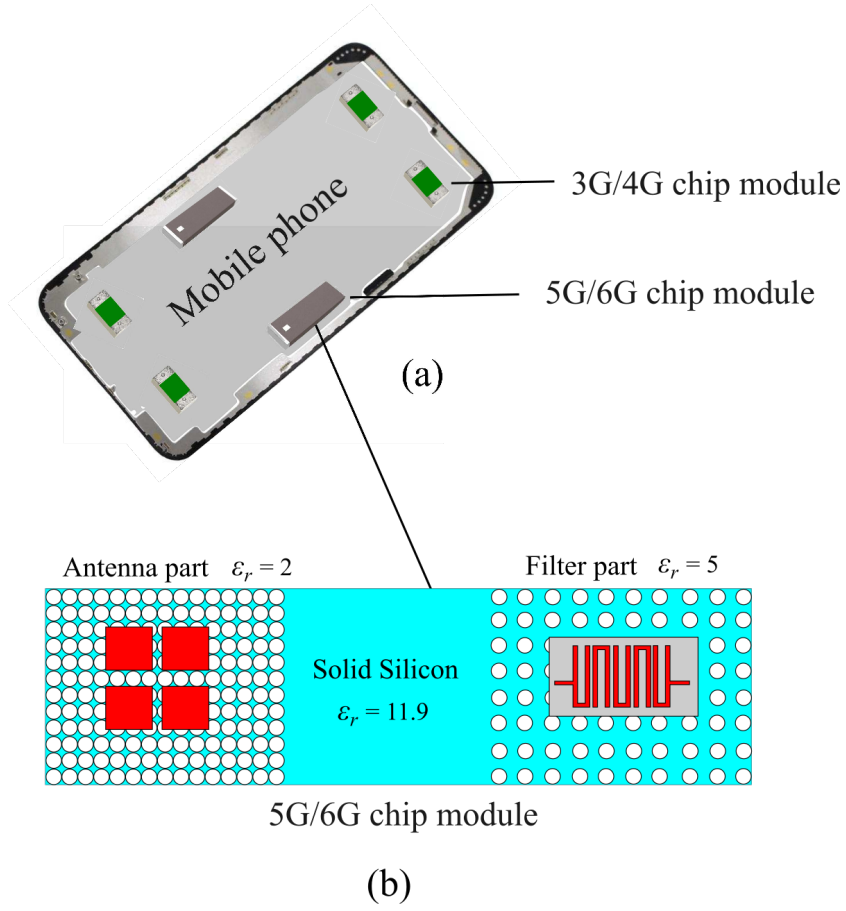


Figure 4.16: Illustration of the proposed ADM in the future mobile phones (a) mobile phone integrated with 3G/4G and silicon-based 5G/6G modules (b) exploded view of the 5G/6G module with ADM having patch antenna array on a lower ϵ_r part and a filter on a high ϵ_r part of the silicon substrate.

calculated as:

$$f_v = \frac{\pi a^2}{4lw} \quad (4.14)$$

It can be seen that by keeping l and w constant, varying a would give different f_v thus controlling the ϵ_{eff} . The real and imaginary parts of the S-parameters are used to extract the constitutive parameters ($\epsilon_{eff}, \tan\delta$) by using the Smith algorithm [140].

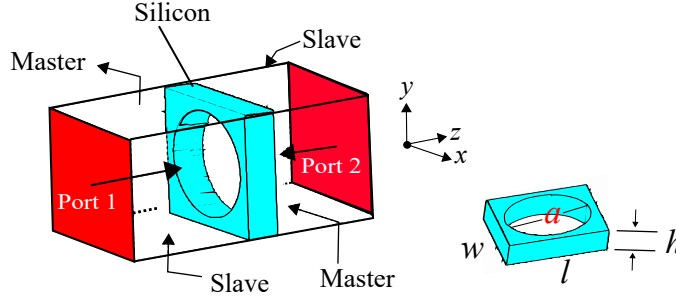


Figure 4.17: Unit cell simulation in the Z-direction for the proposed ADM using Floquet ports with Master and Slave boundaries in HFSSTM.

Figures 4.18 and 4.19 show the ϵ_{eff} and $\tan\delta$, respectively, for the solid Si, porous Si for the filter part and porous Si for the antenna part. For the filter part $a = 70 \mu m$, while for the antenna part $a = 87 \mu m$. It can be seen that for the filter part the $\epsilon_{eff} = 5$ while for the antenna part $\epsilon_{eff} = 2$ which shows a reduction of around 58% and 82%, respectively, compared to the solid Si substrate. It is assumed that these constitutive parameters of Si are constant in this frequency band. Similarly, in the case of the $\tan\delta$, for the filter part it is reduced by 20% while for the antenna part, it is reduced by 55%, compared to the value of solid Si which is 0.015.

4.6.4 Results and Discussion: ADM for Si-Glass based Fabrication

As silicon-based substrates are lossy, there is currently an increasing research trend towards using 3D glass substrates for antennas integrated with other Si-based CMOS components. Glass provides a low-cost and low-loss solution for instance a Borosilicate glass has a ϵ_r of 5.4 and a $\tan\delta$ of 0.005 at 10 GHz [144]. A D-band (130-170 GHz) phased array radio-on-glass (ROG) module was recently

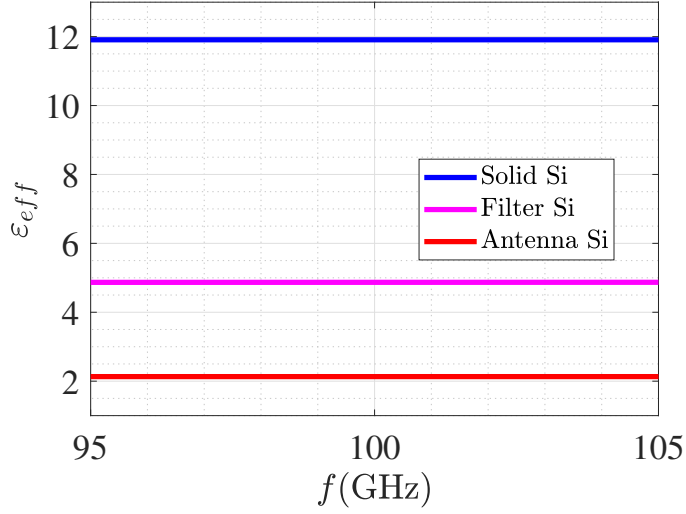


Figure 4.18: Simulated ϵ_{eff} as a function of frequency for solid Si, ADM for filter part when $a = 70 \mu m$ (58% reduction in ϵ_r compared to solid Si), ADM for antenna part when $a = 87 \mu m$ (82% reduction in ϵ_r compared to solid Si).

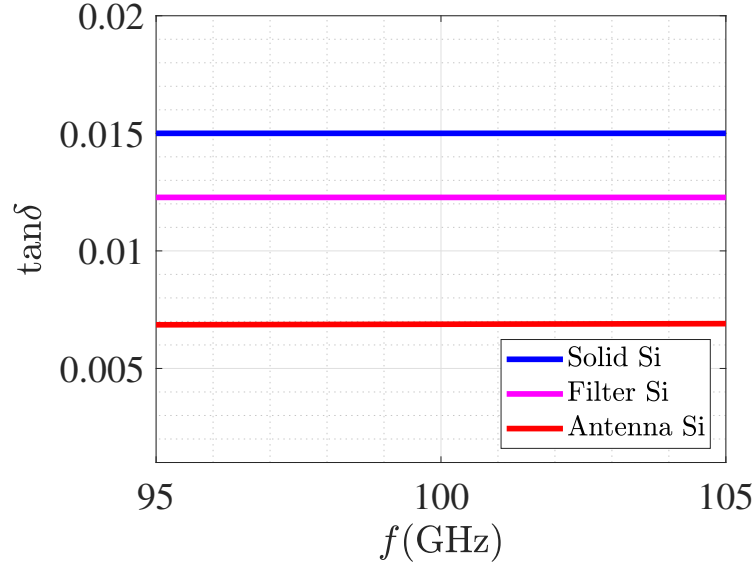


Figure 4.19: Simulated $\tan \delta$ as a function of frequency for solid silicon, ADM for filter part when $a = 70 \mu m$ (20% reduction in $\tan \delta$ compared to solid Si), ADM for antenna part when $a = 87 \mu m$ (55% reduction in $\tan \delta$ compared to solid Si).

designed by Nokia [145]. A unit element is shown in Figure 4.20. The Tx and Rx chips are implemented using a 130 nm SiGe BiCMOS process while the slotted-

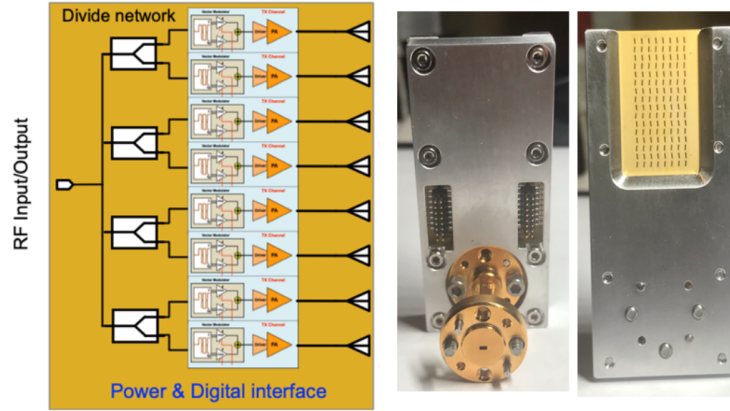


Figure 4.20: D-band phased array Si-glass integrated module. The power divider and chips are on a SiGe substrate connected to a WR-6 waveguide, while the slotted-waveguide array is on a glass substrate [145].

waveguide array is realized on a glass substrate. Figure 4.21 shows another such example at 28 GHz where an LNA-MMIC chip is embedded in a glass cavity along with integrated microstrip patch antennas and band pass filters [146].

Using the proposed technique discussed in the previous section i.e. controlling the f_v , lower values of ε_{eff} and $\tan\delta$ can be obtained for the Si-3D glass integrated solutions. For the illustration purpose, a solid glass substrate with ε_r of 5.4 and a $\tan\delta$ of 0.005 has been loaded with air-cylindrical particles as discussed previously. Simulation results of the ε_{eff} and $\tan\delta$ for the unit cell when $a = 80 \mu m$ are shown in Figures 4.22 and 4.23. The size of the unit cell is $\lambda_g/10 = 87 \mu m$ at 100 GHz while the height is chosen to be $100 \mu m$. From the figures, it can be seen that the ε_{eff} can be reduced from the original value of 5.4 to 2 showing a 62% reduction. Similarly, the $\tan\delta$ can be reduced from 0.005 to 0.00289 showing a reduction of 42%.

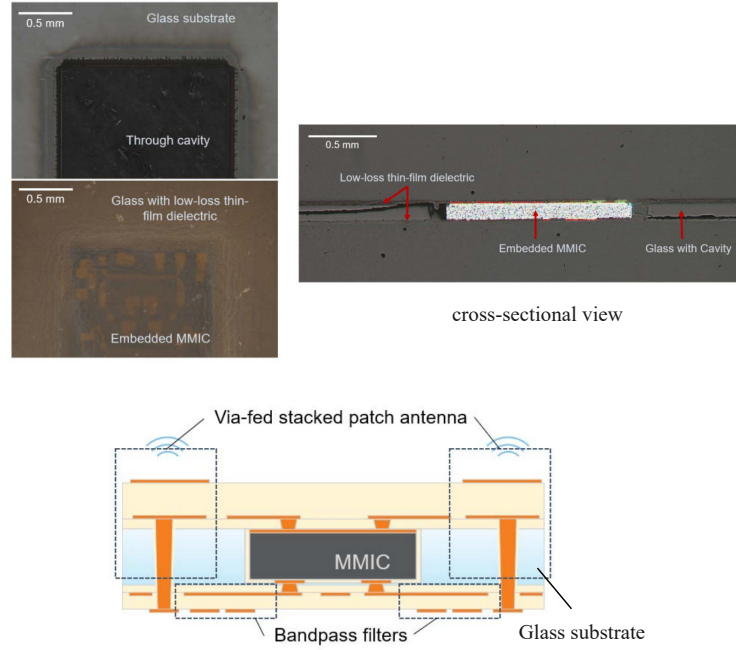


Figure 4.21: Glass-based embedded LNA-MMIC and integrated microstrip patch antenna and filters at 28 GHz [146].

4.6.5 Results and Discussion: Silicon or Glass-based Tune-able Lens Surface for 5G/6G Base Stations

A Lens antenna fed at the focal point transforms the spherical phase front of the feeding antenna into a planar phase front at the opposite side of the aperture with enhanced gain and directivity. They provide an effective solution for beam shaping, low sidelobe levels, and beam steering along with maintaining a compact size. A 3D Luneburg lens is well-known in the literature as having a gradient index profile (the ϵ_r of the lens is varied in a way that the central layer has high ϵ_r while the outer layer has ϵ_r close to 1 ideally) in order to minimize the reflections at each layer.

It is well-known that antennas realized on a lossy silicon substrate have very low gain. In the previous section, it was shown that by using the proposed tech-

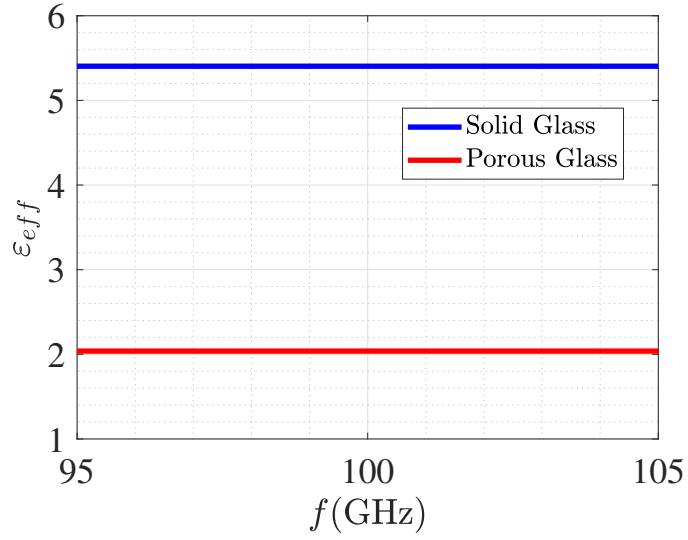


Figure 4.22: Simulated ϵ_{eff} as a function of frequency for the Borosilicate glass substrate when $a = 80 \mu m$ (62% reduction in ϵ_r compared to solid glass).

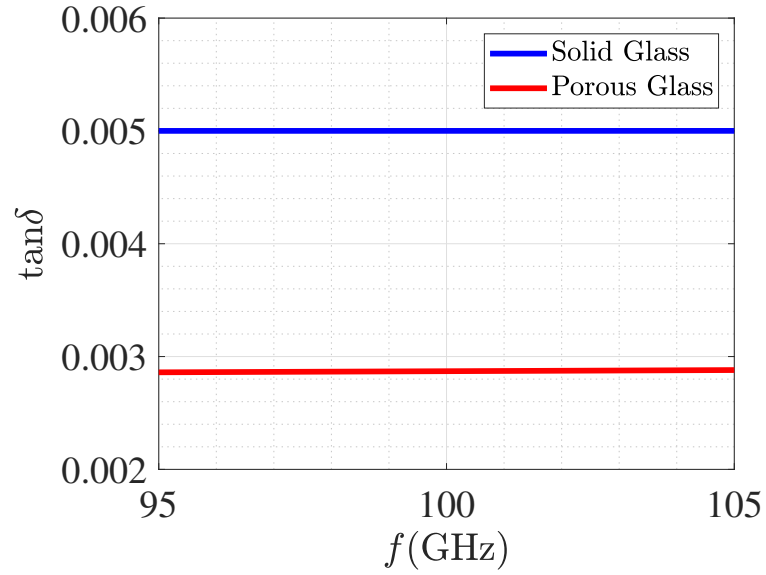


Figure 4.23: Simulated $\tan \delta$ as a function of frequency for the Borosilicate glass substrate when $a = 80 \mu m$ (42% reduction in $\tan \delta$ compared to solid glass).

nique, the silicon substrate can be made less lossy by decreasing its ε_{eff} and $\tan\delta$. Along with using the proposed ADM, a novel idea is proposed that can further enhance the radiation properties of the silicon-based antenna arrays. A Tuneable Lens Surface (TLS) is proposed that can be made of the same material as used for substrates thus no need of additional material is required to implement this TLS making it a low-cost and less complex solution to enhance the gain and to achieve directional radiation patterns.

Figure 4.24 shows the illustration of the proposed lens concept for the semiconductor-integrated antennas to be used in the 5G/6G base stations. Massive MIMO configuration containing hundreds of antenna elements is shown in Figure 4.24(a). The proposed lens surface is based on a 2D gradient concept rather than a traditional 3D spherical Luneburg lens. From Figure 4.24(b), it can be seen that the proposed TLS can be directly laminated on the top of the antenna array. Moreover, the TLS uses the same Si or SiO₂ which is commercially used in semiconductor fabrication processes. In the proposed TLS, the gradient ε_r profile can be obtained by tuning the ε_r of each layer according to the procedure discussed in the previous section. The inner or the center layer can be left solid to have high ε_r or around 12. while the outer layer can be tuned to have a ε_{eff} of 1.1 (close to air) so that the reflections are minimized as shown in Figure 4.24(c). From the analysis presented in the previous section, it was noted that using the perforation technique, the minimum ε_{eff} that can be achieved for the antenna part was 2 reduced from 11.9. We cannot go lower than 2, since the air-cylindrical particle diameter has a maximum dimension that is equal to that of the unit cell. Further increasing the diameter is not possible. For this purpose, to achieve a ε_{eff} of close to 1 for the outer layer of the TLS, a SiO₂ layer can be used for the outer surfaces of the TLS since SiO₂ has a ε_r of 3.9, it can be tuned to have ε_{eff} of close

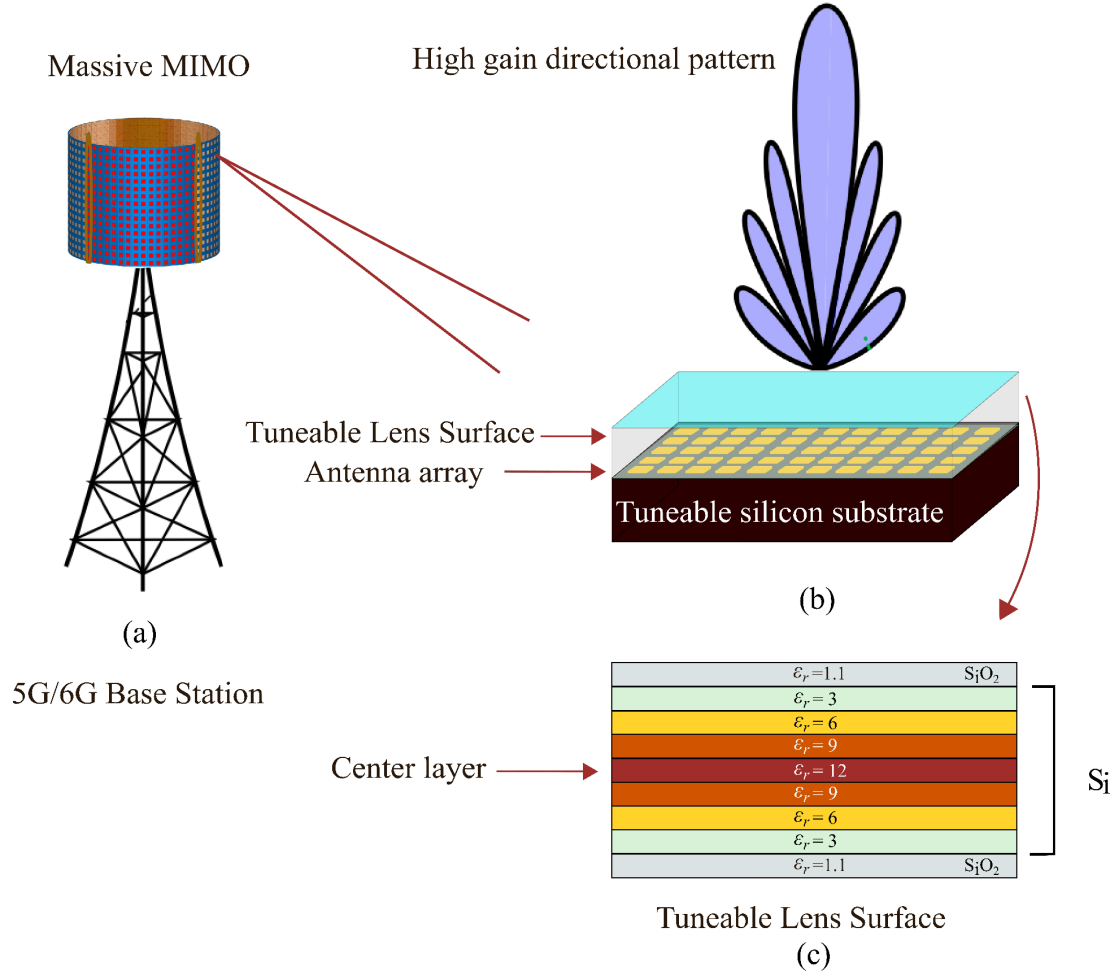


Figure 4.24: Illustration of the proposed TLS integrated on the top of the silicon ADM-based on-chip antenna array (a) 5G/6G base station tower with M-MIMO or UM-MIMO (b) TLS on the top of the on-chip antenna array on a silicon ADM substrate (proposed in the previous section) (c) cross-sectional view of the TLS with gradient ϵ_r profile with a center layer uses the solid Si with roughly $\epsilon_r = 12$.

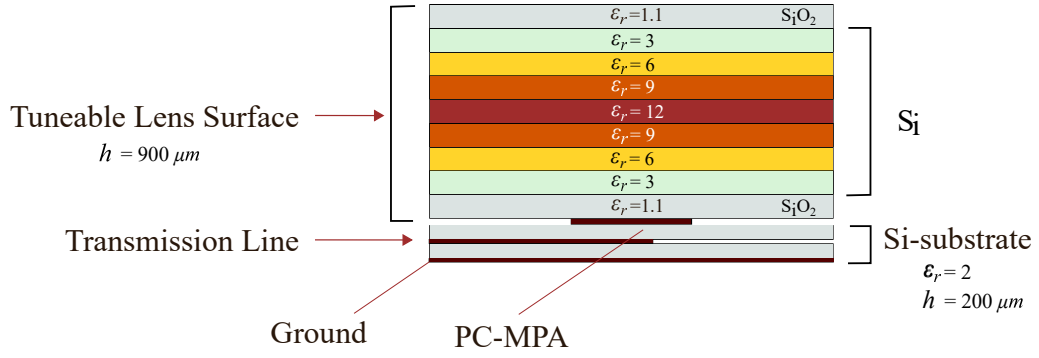


Figure 4.25: Cross-sectional view of the proposed TLS above a PC-MPA in HFSSTM.

to 1.1. It is worth mentioning that, as previously discussed in the earlier section, a glass substrate can also serve as a suitable option for both the substrate and TLS. The proposed TLS structure acting as a superstrate on the top of the microstrip antenna, can be realized by stacking up several artificial dielectric layers of Si and SiO₂ using semiconductor fabrication. The ϵ_{eff} for each layer can be obtained by the particle mixing technique discussed previously.

Figure 4.25 shows the cross-sectional view of the proposed TLS modeled in HFSSTM above the PC-MPA at 200 GHz. The total thickness of the structure is 1100 μm . The substrate has a thickness of 200 μm while each layer of the lens has a thickness of 100 μm with a total of 9 layers. The TLS is designed from a TADL-based Si and SiO₂ material. It should be noted that the structure is not fully optimized at this stage and just initial quick simulation results are shown to prove the concept. Full optimization and analytical modeling can be done in future work.

Figure 4.26 shows the 2D radiation co- and cross-polarization patterns for the PC-MPA at 200 GHz with and without the proposed TLS. The difference in the results is obvious. With TLS (solid lines), the patterns in the E, D, and H planes

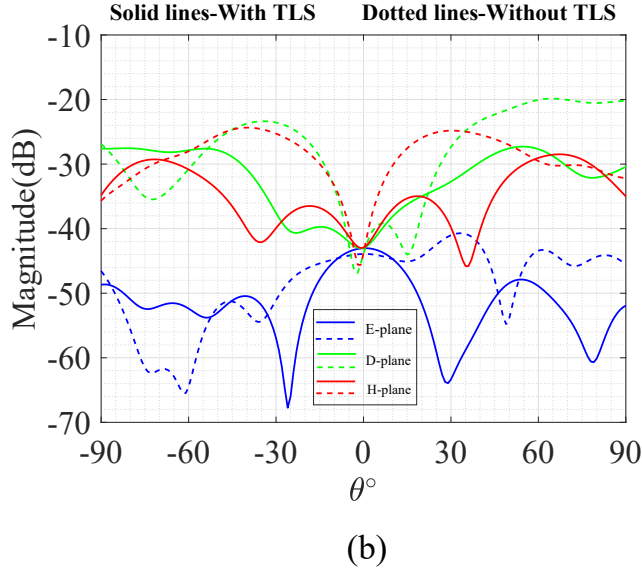
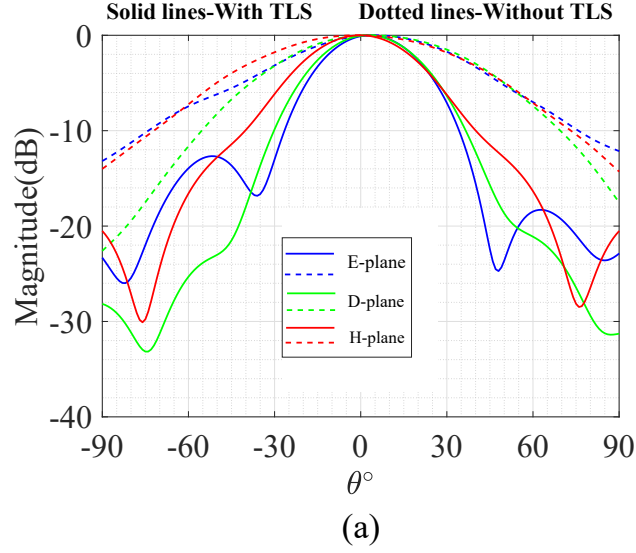


Figure 4.26: 2D co and cross-polarization radiation patterns of the PC-MPA with and without TLS at 200 GHz (a) co-polarized patterns (b) cross-polarized patterns. In the E-plane, the -3 dB beamwidth is now: $\theta_{3dB} = 35^\circ$ compared to 67° showing almost 48% reduction or more directionality. In the D-plane ($\theta_{3dB} = 37^\circ$ compared to 69°) and H-plane ($\theta_{3dB} = 40^\circ$ compared to 78°) the reductions are 46% and 49%, respectively.

are now directional compared to the non-TLS case (dotted lines). The gain has been improved by 2 dB. In the E-plane, the -3 dB beamwidth is now: $\theta_{3dB} = 35^\circ$ compared to 67° showing almost 48% reduction or more directionality. Similarly, in the D-plane ($\theta_{3dB} = 37^\circ$ compared to 69°) and H-plane ($\theta_{3dB} = 40^\circ$ compared to 78°) the reductions are 46% and 49%, respectively. Figure 4.26(b) also shows improvements in the cross-polarized patterns for the TLS case compared to the non-TLS case.

4.7 Summary

In this chapter, theory, procedures, and analysis for designing artificial dielectric material based on the Maxwell Garnett theory have been thoroughly discussed. Correct analytical models of Maxwell Garnett mixing theory were used to obtain a high-performance RF material, from an existing highly thermally stable and compatible material. A complete set of design equations for finding desirable ε_{eff} of both isotropic and anisotropic particles was presented. The permittivity analysis for both isotropic and anisotropic cases was thoroughly studied including the effects of the direction of excitation as well as the polarization of the incident field on the ε_{eff} . Detailed design procedures and requirements were discussed for both microwave and mm-Wave applications. Several analytical models were discussed and evaluated for their performance. The most accurate model was identified and used to find the ε_{eff} of the anisotropic cylindrical particle in any direction of excitation as well as any polarization of the incident field. Building on all the advantages of the existing substrate materials, the proposed method can be used to obtain the desired lower ε_r and $\tan\delta$ values. Thus, enabling the use of inexpensive materials to be used as high-performance RF substrates for applications that

require high antenna efficiency, multi-layer PCB compatibility, and high thermal stability. For the proof of concept, several samples were fabricated and tested in the S-band at 3.5 GHz as well as in the W-band at 77 GHz. A reduction of 45% in the ε_{eff} and 38% in the $\tan(\delta)$ values were obtained in the S-band, and 32% and 72% reduction in ε_{eff} and $\tan\delta$, respectively, in the W-band. Excellent agreement was obtained between the simulation, calculated, and measured results.

It was also shown that the proposed particle mixing technique has the potential to be used in semiconductor fabrication at THz frequencies to reduce the losses associated with the substrate having a high ε_r for on-chip antennas. Finally, a TLS structure was also discussed to enhance the gain and improve the radiation performance of on-chip antennas at THz frequencies. In the following chapter, various solutions, including the artificial dielectric lens based on the particle mixing theory, are discussed to address the wide beamwidth problem associated with QRHAs for UAV-based radar characterization.

Chapter 5

Ultrawideband Artificial Dielectric Lens-Integrated Antenna Solutions

5.1 Introduction

Ridged horn antennas seem a promising antenna choice for a variety of UWB radar and communication system applications. However, in Chapter 2, it was mentioned that due to the wide beamwidth problem of these ridged horn antennas, the radiation performance of the antenna probe is adversely affected in terms of the mismatch and cross-polarization, when the UAV platform is taken into consideration. Chapter ?? then established the theory and foundation work for designing artificial dielectric materials at both microwave and mm-Wave frequencies. This chapter uses this theory and principles in the design of a 3D-printed graded index (GRIN) artificial dielectric lens antenna (ADLA), thus (GRIN-ADLA), presenting an interesting and feasible solution to the wide beamwidth problem of commercial ridged horn antennas for UAV-based characterization of radars. Towards the end of the chapter, several other lenses as well as corrugated ridged horn antenna solutions are also briefly presented as suitable configurations for beamwidth reduction problems in these commercial probe antennas.

5.2 Dielectric Lenses: Background

Because of the growing number of applications, availability of low-loss dielectric materials, and advanced fabrication techniques, the interest in dielectric lenses has been renewed in the research community.

A conventional lens can take any structure and most commonly: hyperbolical, bi-hyperbolical, elliptical, or hemispherical lenses are reported in the literature. Such structures are mainly used to converge the scattered energy coming from the feed to make a directive beam [147]. Lens antennas are end-fire radiators having a single directive radiation beam in one specific direction. From the literature, it can be noticed that the dielectric lens antenna can be constructed from conventional dielectrics such as Lucite or polystyrene and also from artificial dielectrics.

Lens antennas transform spherical phase fronts into planar phase fronts across an aperture to enhance their directivity. Lens antennas provide an effective solution for beam shaping, low sidelobe levels, and beam steering along with maintaining a compact size. Dielectric Lenses are more advantageous as they can be efficiently constructed using simple molding or 3D printing.

Lens antennas provide flexibility in terms of feeding. In literature works, horn antennas, dipoles, microstrip patch antennas, and even arrays of antenna elements are used to feed the lens structure. However, open-ended waveguides and horn antennae are widely used for feeding the lens.

A spherical lens having a uniform dielectric material inside is termed a “constant-index lens”. It has a mediocre performance in terms of collimation of the feed energy and sidelobe level performance. However, a “variable-index” or “spherical stepped-index” lens also well-known as a Luneburg lens [148] is usually used to achieve optimum performance. An illustration of rays emitting from a

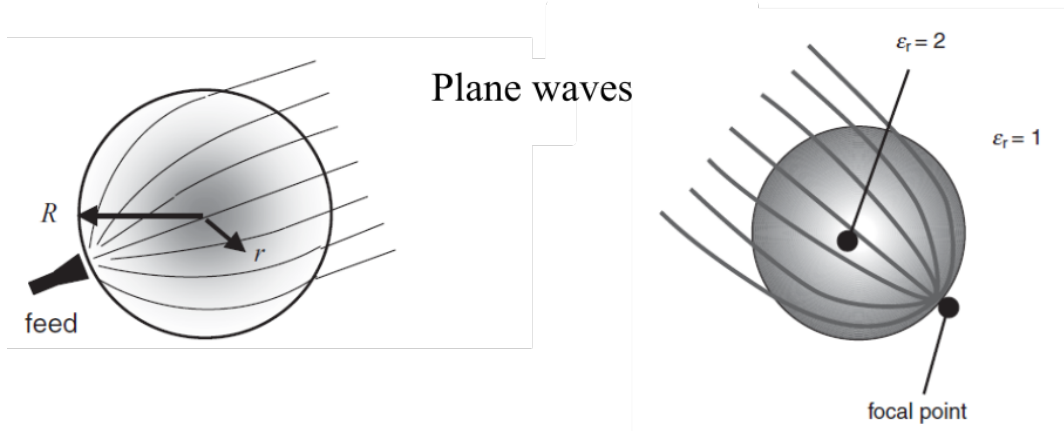


Figure 5.1: Ray paths showing beam collimation in a typical spherical stepped-index Luneburg lens [147].

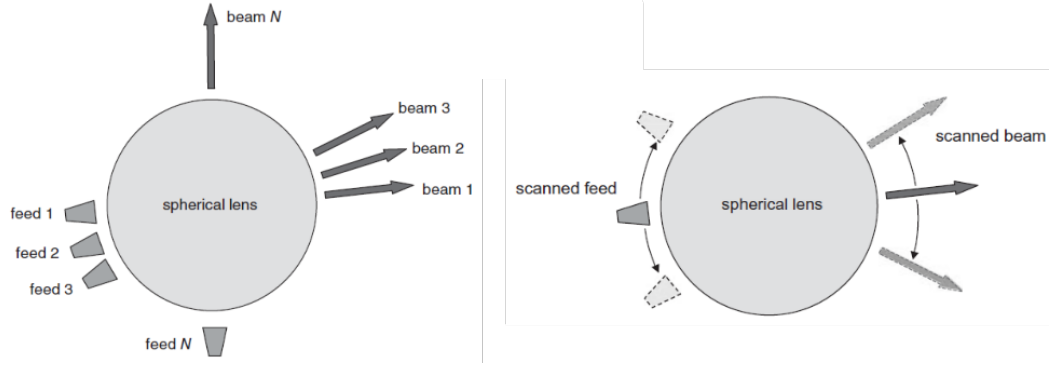


Figure 5.2: Illustration of beam scanning using a spherical lens antenna with multiple feeds [147].

typical Luneburg lens forming beam collimation is shown in Figure 5.1 [147]. In such lenses, the permittivity varies according to (5.1) given below:

$$\epsilon_r = 2 - \left(\frac{r}{R}\right)^2 \quad (5.1)$$

where r is the axial radius from the center point and R is the outer radius of the lens.

The Luneburg lens provides excellent performance but at the cost of the com-

plexity of the design due to the gradient refractive index profile. Such a structure having multiple concentric dielectric layers, is difficult to fabricate, especially at mm-Wave frequencies. However, simpler lenses having constant index or two-layer lenses consisting of two different permittivity values can be easily fabricated and it also ensures good performance [149].

Spherical lens antennas are simple in configuration and have good electrical performance. They can provide wide-angle beam scanning without losses. If the spherical lens is fed with an array of feeding antennas, beam scanning can be digitally achieved by switching from one feed to another feed or through mechanical rotation of the feeds around the lens [147]. Figure 5.2 demonstrates the beam scanning mechanism by using spherical lenses having multiple feeds.

An approximate estimation of the HPBW from a circular aperture is given by (5.2) [150]:

$$\theta_{3dB} = \frac{29 \times \lambda}{R} \quad (5.2)$$

Using (5.1) and (5.2), the dielectric profile of a Luneburg lens can be plotted when the beamwidth is known. Figure 5.3 shows the dielectric profile of this lens for different HPBW values with respect to the normalized radius of the lens. For this profile, the frequency is assumed to be 84 GHz. It can be seen that for the smaller HPBW values, the lens's overall radius R is larger since the HPBW is inversely related to the size of the antenna aperture.

The focal point (f) of a typical spherical Luneburg lens is the outer surface. However, in some cases, f can be modified based on the source antenna used to excite the lens structure. A modified permittivity profile of a Luneburg lens in

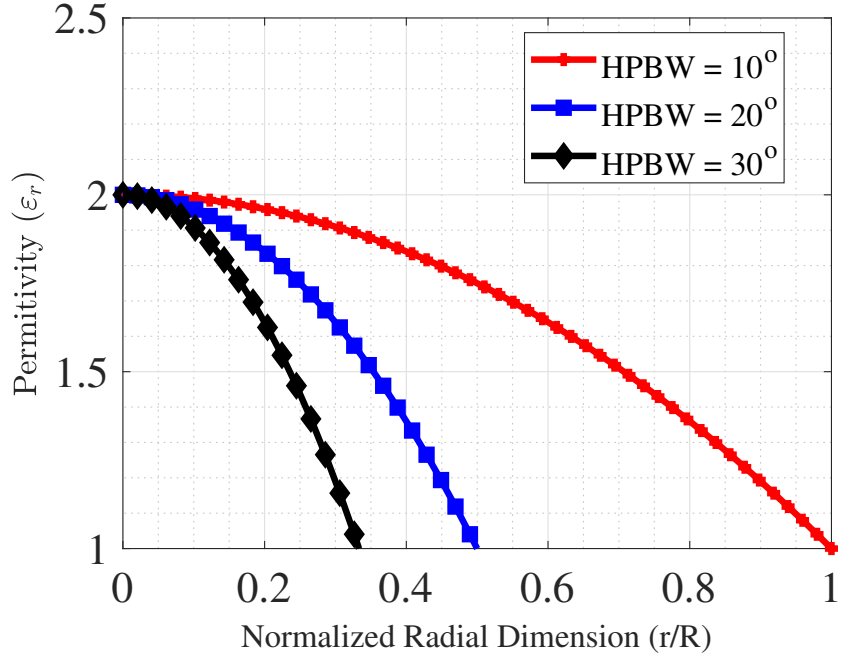


Figure 5.3: Permittivity profile of a typical Luneburg lens.

which f can be adjusted is given by (5.3) [151] and shown in Figures 5.4 and 5.5. For this profile, the frequency is assumed to be 77 GHz while the HPBW is chosen to be 10° .

$$\epsilon_r = \frac{1 + f^2 - (\frac{r}{R})^2}{f} \quad (5.3)$$

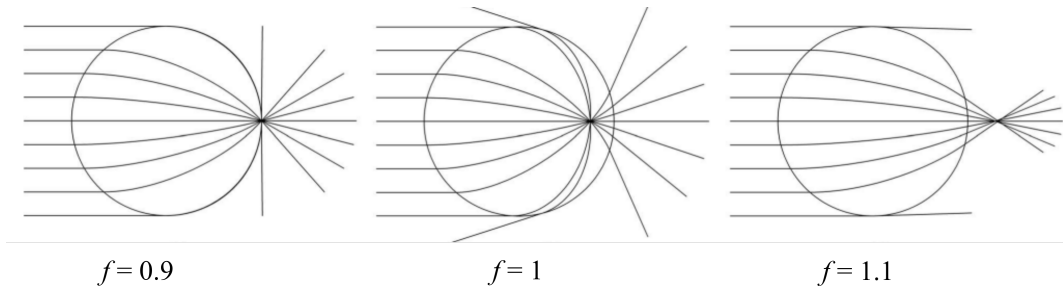


Figure 5.4: Ray tracing trajectory in a modified Luneburg lens [151].

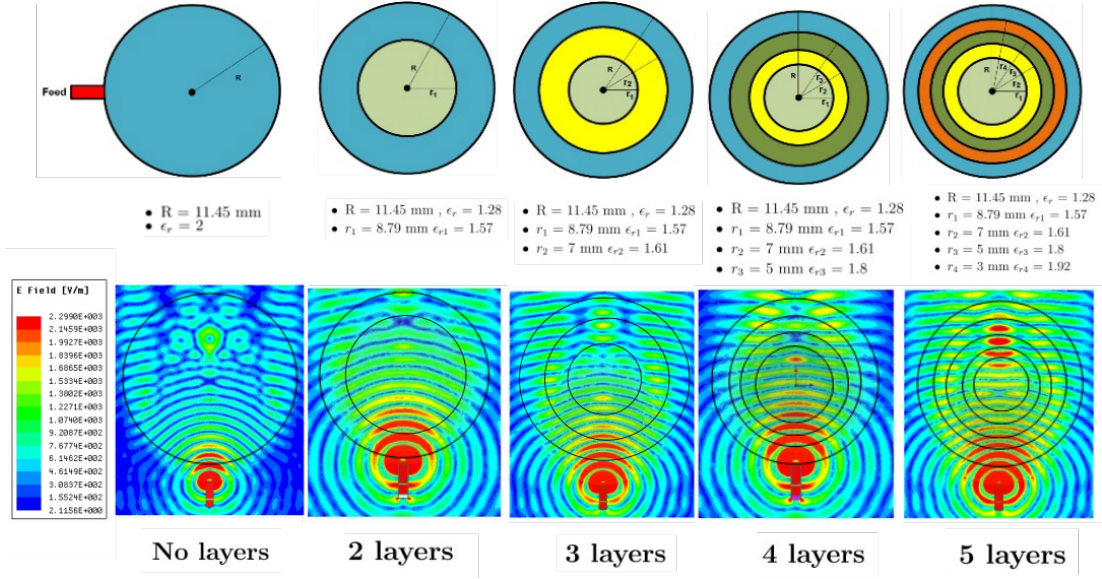


Figure 5.6: Geometry and E-field (V/m) distribution of the multi-layer Luneburg lens.

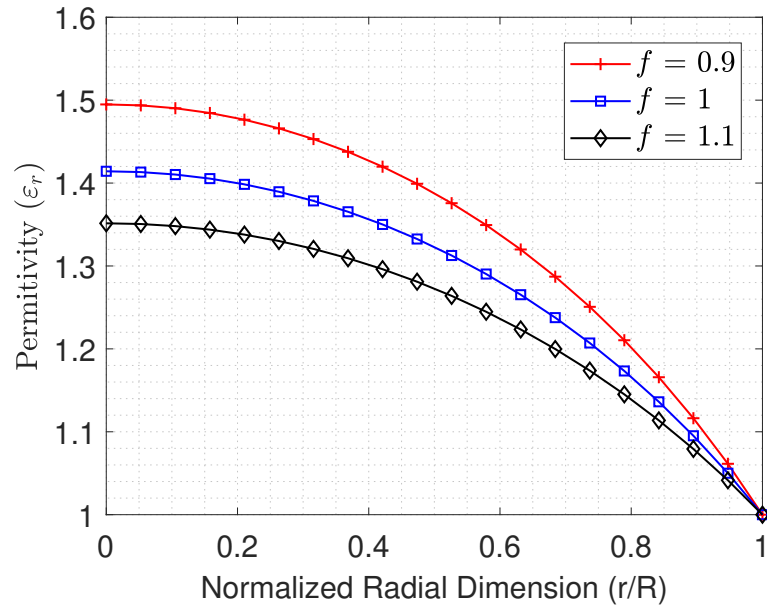


Figure 5.5: Permittivity profile of a modified Luneburg lens

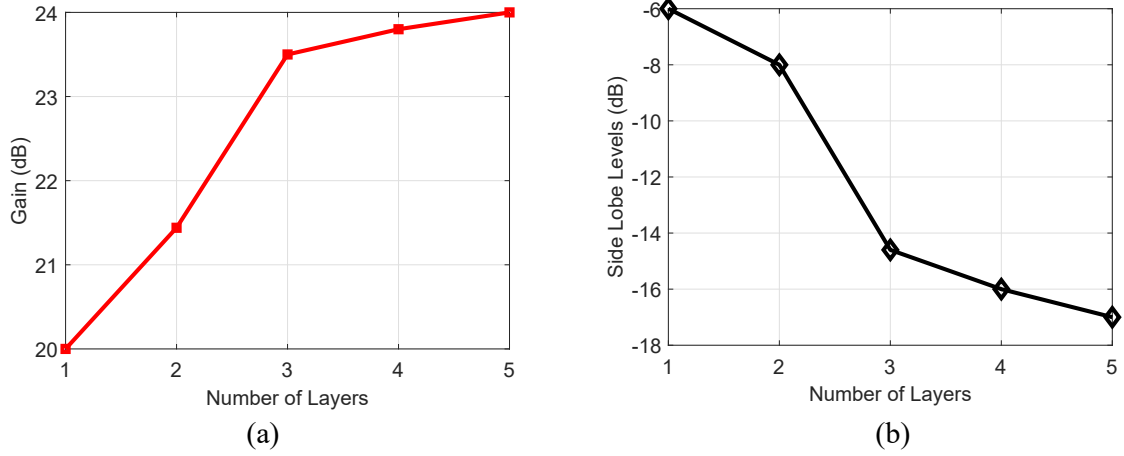


Figure 5.7: Effect of the number of layers on the gain and sidelobe level performance (a) Gain (b) sidelobe level.

5.3 A Spherical Luneburg Lens: Results and Discussion:

To validate the aforementioned Luneburg lens theory, a spherical Luneburg lens with 5 gradient layers is designed and the results are presented. The geometry and E-field (V/m) distribution inside the lens are shown in Figure 5.6. The antenna is designed at 77 GHz, with a total size $R = 22.90$ mm in order to have a HPBW of around 10° . A rectangular waveguide is used to feed this antenna on the surface. Both the lens and the source antennas are modeled and simulated in HFSSTM. One can observe that as the number of layers increases, there is a noticeable improvement in field collimation, resulting in more directed energy compared to scenarios with no layers. This highlights the effectiveness of the Luneburg lens. Figure 5.7 shows the effect of the number of layers on the gain and sidelobe level performance. It can be seen that with increasing the number of layers and thus due to small discretization of permittivity (minimal internal reflection within layers), the gain is increased and the sidelobe level is reduced. Figure 5.8 shows the normalized radiation patterns in both E- and H-planes of the Luneburg lens.

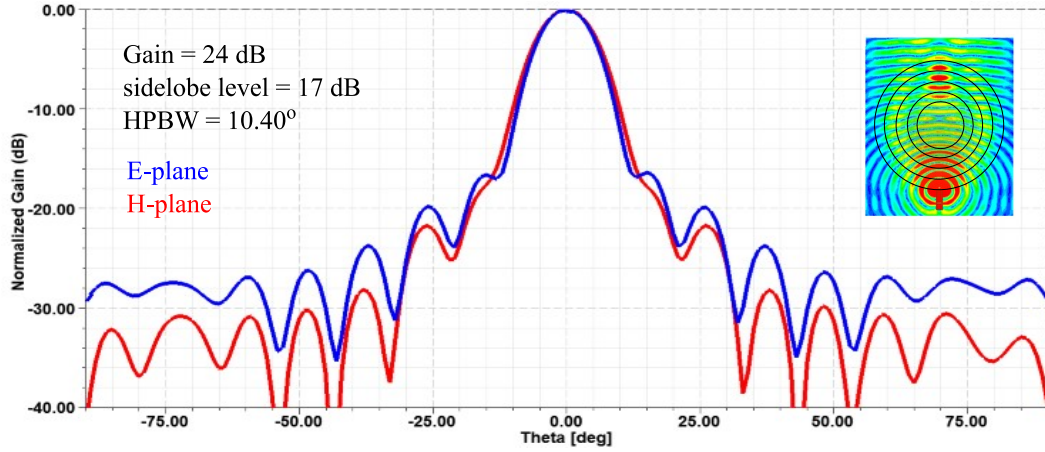


Figure 5.8: Normalized 2D radiation patterns of the multi-layer Luneburg lens.

It has a gain of 24 dB, sidelobe level of -17 dB, and a HPBW of 10.40° .

For the purpose of demonstration, only a uniform dielectric lens is fabricated using a Formlabs Form 3 resin-based 3D printer, since it is challenging to fabricate such a multi-layer lens with varying dielectric profile with the available 3D-printed fabrication capabilities. Figure 5.9 shows the 3D printed prototype of the lens along with its mm-Wave RF scanner near-field measurement setup. Figure 5.10 shows the comparison of the simulated and measured 2D radiation patterns of the lens in the E-plane revealing a high degree of agreement in the results. A slight mismatch at 15° may be attributed to non-uniformity resulting from the manual polishing of the printed lens or could be due to a slight misalignment in the measurement setup.

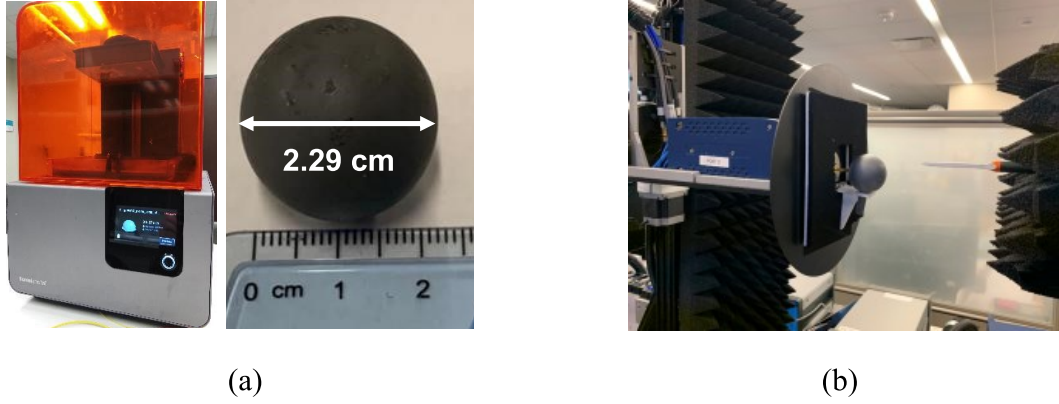


Figure 5.9: Uniform dielectric lens fabricated prototype (a) 3D printed lens using Formlabs Form 3 resin-based 3D printer (b) near-field RF scanner measurement setup .

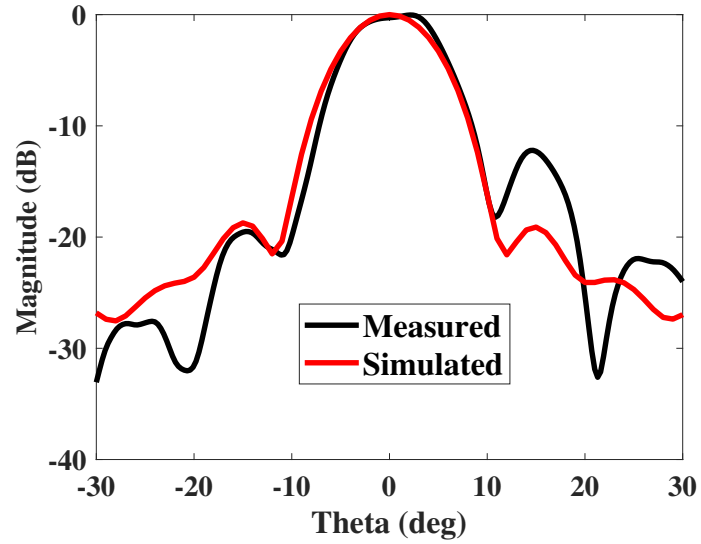


Figure 5.10: Comparison of the simulated and measured 2D patterns of the lens in the E-plane.

5.4 Planar UWB Graded-Index Lens

Several works can be found in the literature related to beamwidth reduction and beamwidth constancy in ridged-horn antennas. These beamwidth constancy or pattern stability techniques can be divided into three main categories i.e. using

H-plane pinwalls [152], shaped ridges [153] or shaped sidewalls [154], and corrugations [21, 155–159]. In [152], beamwidth constancy has been achieved over a band of 4–16 GHz, by combining two techniques: using pinwalls in the H-plane and shaped walls in the E-plane. However, the horn has a large length of around 9λ at the lowest frequency of operation which is around 635 mm. Moreover, the design has significant pattern degradation at higher frequencies of the band. The design presented in [153] has combined the technique of using pin walls in the H-plane with the shaped ridges in the E-plane, however, limited contribution has been made to achieve beamwidth constancy as well as pattern degradation over the entire band of operation. The design reported in [154] has used a conical horn with shaped boundary walls to achieve a constant 10 dB beamwidth. However, the design has a narrow bandwidth of around 2.25 GHz. The corrugated designs presented in [21, 155–158] have either narrow bandwidth (corrugated slots have dimensions in terms of wavelength), targeted high frequencies: 8–18 GHz, or have used complex geometries. Fabrication of such designs usually involves four-axis CNC machining which increases the overall cost of the design.

The availability of a range of low-loss dielectric materials, coupled with the development of low-cost, high-precision 3D-printing technology, has facilitated the successful implementation of numerous high-performance planar graded index (GRIN) lens antennas. Several works related to 3D-printed planar GRIN lens antennas can be found in the literature [160–164]. Dielectric or metamaterial-based lens antennas have been used with horns or ridged-horns to reduce the phase error and to improve their radiation performance [165–171]. However, these works have not focused on beamwidth constancy, pattern symmetry, and also have narrow frequency bandwidth performance.

The proposed GRIN-ADLA is designed based on the previously established

Maxwell Garnett particle mixing theory. This antenna provides a novel, cost-effective, simple, and efficient wideband solution to existent antennas that has asymmetric radiation pattern and wide HPBW problems. The wideband performance of the proposed GRIN-ADLA broadens the range of applications by covering several maritime, weather, surveillance, and airborne radar frequency bands. The incorporation of the proposed GRIN-ADLA allows for the effective utilization of commercially available wideband probe antennas (SH2000 and QH2000) on UAVs, thus addressing a previous limitation caused by the undesirable back-scattering resulting from the wide beamwidth performance of these commercial probes.

The proposed GRIN-ADLA ensures achieving symmetric radiation patterns as well as reducing the wider beamwidth performance in the E- and H-plane patterns from $120^\circ\text{--}40^\circ$ to $35^\circ\text{--}20^\circ$ within the desired 3–18 GHz frequency band, providing an average beamwidth reduction of 68%, thus satisfying the probe antenna requirements set for conducting radar characterization using UAVs in the field. The required values for half-power beamwidth and co-polar pattern mismatch are less than 40 degrees and 0.2 dB, respectively, and the proposed probe antenna meets these requirements. The proposed antenna is fully characterized and tested in the frequency range of 3–18 GHz, and the simulation, as well as measurement results, show a very good agreement, confirming the suitability of the proposed antenna system for in-situ antenna measurements of radars and communication systems. To the best of the author’s knowledge, the proposed work is the first of its kind in the open literature that uses a simple 3D-printed GRIN lens antenna to achieve pattern symmetry, beamwidth reduction, and beamwidth constancy in such a wide frequency band of operation. These features make this antenna an ideal candidate for not only UAV-based radar characterization applications but

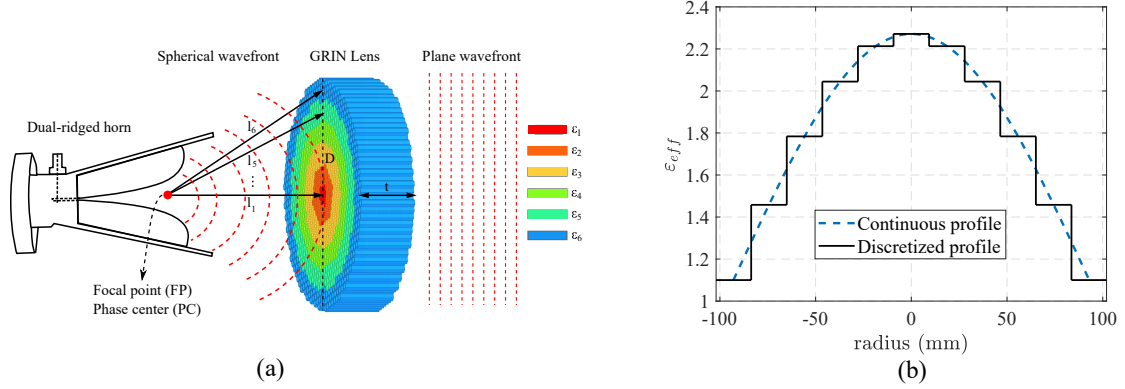


Figure 5.11: Antenna geometry models used in this work (a) illustration of the plane wavefront obtained using the proposed GRIN lens excited by SH2000 feed antenna (b) graded ϵ_{eff} profile of the GRIN lens.

also open the door for a range of other UWB communication applications where “well-matched” patterns are required such as in EMC testing systems.

5.4.1 Design Procedure and Fabrication

5.4.1.1 Lens Design and Formulation

Figure 5.11(a) shows the geometry model of the proposed GRIN lens consisting of six rings having a gradient distribution of ϵ_{eff} across the aperture. Where D is the diameter of the lens, t represents thickness, ϵ_1 is the ϵ_{eff} of the first ring, and l_1 is the distance of the first ring from the focal point (FP) of the lens. The lens is excited by the commercially available dual-ridged horn antenna [110].

By approximating the desired radiation pattern coming out of a circular aperture, the diameter (D) of the aperture can be related to the desired HPBW at the output of the aperture (θ_{3dB-o}) in degrees, by the expression [172]:

$$D = \frac{58.5\lambda_o}{\theta_{3dB-o} \times \varepsilon_T} \quad (5.4)$$

where ε_T is taper efficiency, and for a circular aperture, $\varepsilon_T \approx 0.85$, considering a non-uniform illumination of the aperture [172]. In the above expression λ_o is the free-space wavelength and its value is 100 mm considering the lens is designed at the lowest frequency of the band i.e. at 3 GHz. Using the desired HBPW value of 35° at 3 GHz, the value of $D \approx 197$ mm. The FP can then be determined by using simple trigonometry and is given by the expression:

$$FP = \frac{D}{2 \tan(\theta_{3dB-s}/2)} \quad (5.5)$$

where θ_{3dB-s} is the HPBW of the source antenna used to illuminate the lens.

In order to achieve the required phase shift and desired planar wave propagation at the output of the lens, a dielectric constant profile is required for the lens. For the gradient permittivity profile, a similar procedure reported in [160] was followed to obtain the desired profile using the expression presented in (5.6). In consideration of the fabrication tolerances and the need for a sturdy structure, the ε_{eff} for the outermost ring ε_6 was established at 1.1. This corresponds to a relatively lower achievable ε_{eff} , slightly higher than that of air.

$$\varepsilon_1 = \left[\frac{l_6 - l_1 + \sqrt{\varepsilon_{r6}}t}{t} \right]^2 \quad (5.6)$$

where t is the thickness of the lens, ε_1 is the ε_{eff} of the innermost ring, and

l_1 and l_6 are the respective distances of the corresponding rings from the focal point. From (5.6), it can be observed that the thickness of the lens dictates the gradient permittivity profile, and hence, the thickness is usually assumed to have a certain value that provides the desired achievable ε_r provided the values of l_1 or FP , l_6 , and ε_6 are known [160]. The lens thickness was selected as 50 mm in order to obtain an achievable profile, given that the ε_r of the material is 2.7.

To calculate FP or l_1 , using (5.5), the HPBW of the source antenna is required. Due to the asymmetric patterns, the HPBW of the dual-ridged horn antenna at 3 GHz has two different values of 118° and 58° in the H and E plane, respectively [110]. It can be seen that when $\theta_{3dB-s} = 118^\circ$, $FP = l_1 = 59.18$ mm. However, when this value is used in (5.6), gives $\varepsilon_1 = 4.6$, which is greater than the maximum dielectric value of the core material 2.7. Therefore, the other value of the beamwidth $\theta_{3dB-s} = 58^\circ$ was used that makes $FP = l_1 = 177$ mm. Interestingly, it can be seen that in order to fully illuminate the desired aperture, the focal point is adjusted based on the beamwidth of the source. For instance, when the beamwidth of the source is wider ($\theta_{3dB-s} = 118^\circ$), the source antenna is placed near to the aperture ($FP = l_1 = 59.18$ mm), and when the beamwidth is narrow ($\theta_{3dB-s} = 58^\circ$), the source antenna is placed at a far distance from the aperture $FP = l_1 = 177$ mm, in order to have efficient illumination of the lens. Figure 5.11(b) shows the gradient permittivity profile calculated based on (5.6) when $FP = l_1 = 177$ mm, $l_6 = 118$ mm (using trigonometry), and $t = 50$ mm.

5.4.1.2 Unit Cell Design and Formulation

Hexagonal unit cells for each ring were first designed and characterized in HFSSTM using periodic boundary conditions and floquet ports as shown in Figure 5.12. The core material was loaded with air-cylindrical particles using the theory of

Maxwell-Garnett discussed in detail in Chapter ?? as well as in [124], and the f was varied by varying the radius r of the cylindrical particle, to achieve desired ε_{eff} (ε_{eff}) for each ring of the proposed lens. This procedure is also previously discussed in Chapter ?? and also in [19]. The cylindrical particle was chosen due to its simplicity and ease of fabrication. A commonly available Formlabs Form 3 3D printer having photopolymer general purpose grey resin (RS-F2-GPGR-04) was used as a core material having a ε_r of around 2.7 at 3 GHz. The f of the cylindrical particle inside a hexagonal unit cell is given by the expression:

$$f = \frac{2\pi r^2}{3\sqrt{3}s^2} \quad (5.7)$$

where r represents the radius of the air-cylindrical particle, and s is the length of one of the sides of the hexagon as shown in Figure 5.12. As discussed in [19], ε_{eff} of an anisotropic particle is dependent upon the direction of the propagation of the wave. In this case, as the wave propagation is parallel to the length or the t -dimension of the cylinder, Van Beek model [134] represented by the expression in (5.8) was used to find to ε_{eff} in the desired direction.

$$\varepsilon_{eff} = \varepsilon_b + \frac{f}{3} \frac{(\varepsilon_a - \varepsilon_b)(5\varepsilon_b + \varepsilon_a)}{(\varepsilon_b + \varepsilon_a)} \quad (5.8)$$

where ε_a is dielectric constant of the loaded particle (in the case of air-cylindrical particle, $\varepsilon_1 = 1.0006$), and $\varepsilon_b = 2.7$ for the core Resin material. The ε_{eff} values obtained from the HFSSTM were verified using Van Beek analytical expression of (5.8). For brevity, the comparison of ε_{eff} values from simulations and the analytical model is not shown here but can be found in detail in [19].

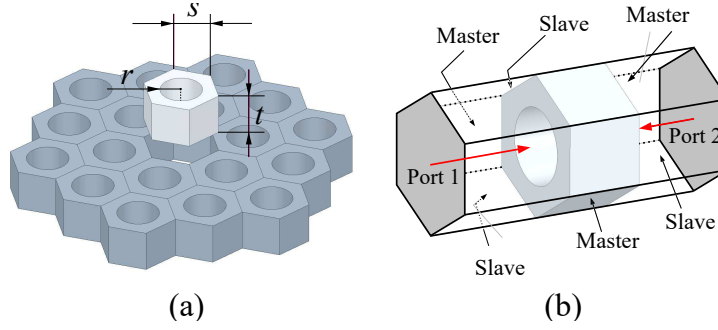


Figure 5.12: Hexagonal Unit cell used to build the proposed GRIN lens (a) description of the unit cell and its dimensions, where $s = 3.67$ mm ($0.06\lambda_g$), and $\lambda_g = \lambda_o/\sqrt{\epsilon_r}$, $t = 50$ mm ($0.82\lambda_g$), and r is varying radius (b) unit cell simulation setup in HFSSTM using Floquet ports and periodic boundaries.

Table 5.1: Calculated parameters of the proposed GRIN lens.

Ring	Dielectric constant	Fractional volume	Hole radius (mm)
1	2.2713	0.1931	1.4667
2	2.2130	0.2194	1.5632
3	2.0445	0.2953	1.8136
4	1.7838	0.4128	2.1442
5	1.4580	0.5595	2.4965
6	1.1000	0.7208	2.8336

To achieve the desired values of ϵ_{eff} for each ring of the lens, the side of the hexagonal unit cell was kept constant to $s = 3.67$ mm for all the rings. This allowed for the variation of r based on the fabrication tolerances as well as covering the whole lens aperture without any gaps.

Following the dielectric constant profile shown in Figure 5.11(b), r was varied to obtain different values of f using (5.7) for each ring of the lens. These values of f were then used to calculate ϵ_{eff} using (5.8). Table 5.1 summarizes these values for the proposed lens. The complete lens was finally modeled in HFSSTM and then fabricated in four parts which were assembled together using epoxy. These 3D-printed prototypes are shown in Figure 5.13.

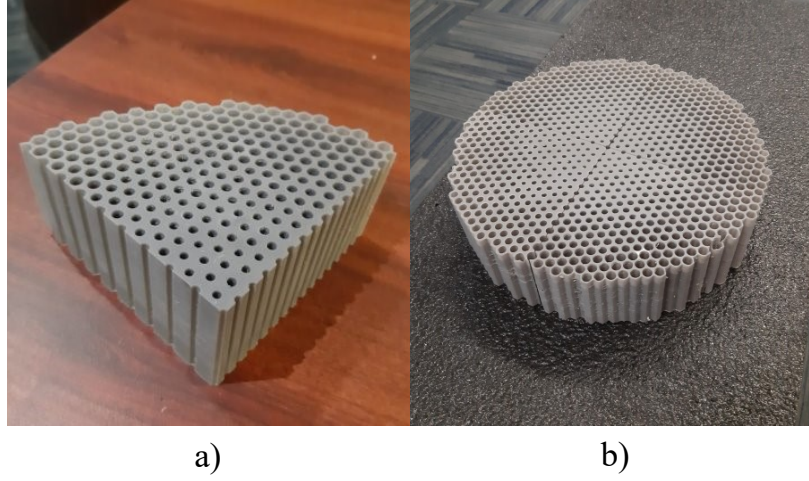


Figure 5.13: 3D-printed GRIN lens prototypes a) quarter section of the lens b) complete assembled lens.

5.4.2 Supporting Frame: Design and Analysis

A detailed analysis was conducted before selecting a suitable frame to hold the proposed GRIN lens structure. Several parameters were investigated including the robustness of the frame and its effect on the radiation as well as the S-parameter performance of the antenna. Two frame designs as shown in Figure 5.14 were modeled and analyzed for their performance in HFSSTM. Figure 5.14(a) shows the complete conical model encapsulating the whole structure. The frame was 3D-printed using ABS plastic material as shown in Figure 5.14(b). For the second frame model depicted in Figure 5.14(c), two Rohacell® material: Rohacell® 300WF and Rohacell® 71HF, were utilized, and the frame was subsequently CNC-machined using Rohacell® 300WF shown in Figure 5.14(d).

Figure 5.15 shows the effect of these frames on the radiation and S-parameter performance of the lens antenna. Both co and cross-polarization patterns were compared for various cases. Figures 5.15 (a) and (b) show the radiation patterns in the E- and H-planes, respectively, while Figure 5.15(c) shows the cross-polarization

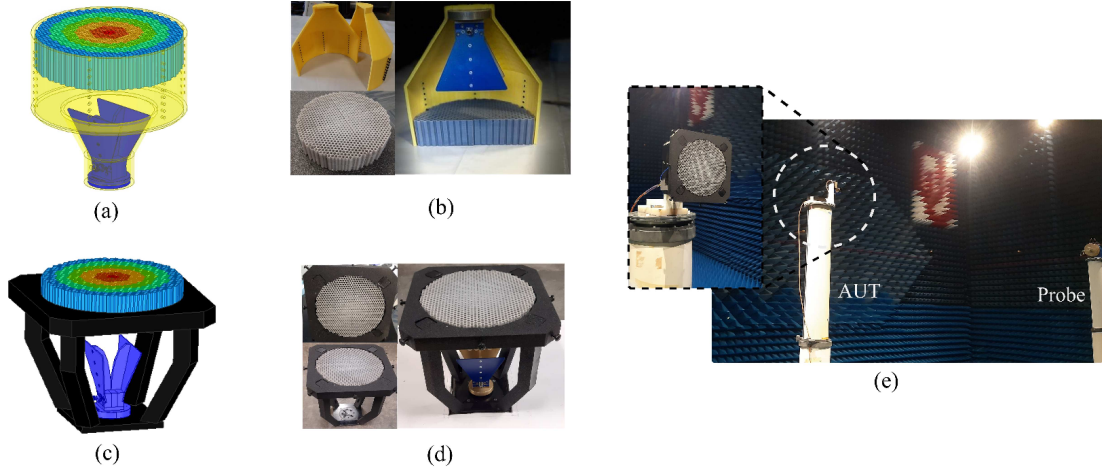


Figure 5.14: Simulated and fabricated 3D frame models for holding the lens (a) simulated model with ABS plastic frame ($\epsilon_r = 2.75$, $\tan\delta = 0.025$) (b) 3D-printed ABS frame model (c) simulated model with Rohacell® 300WF ($\epsilon_r = 1.3$, $\tan\delta = 0.009$) and Rohacell® 71HF ($\epsilon_r = 1.093$, $\tan\delta = 0.0038$) frames (d) CNC-machined fabricated Rohacell® 300WF frame model (e) far-field anechoic chamber measurement setup.

patterns. From Figures 5.15 (a) and (b), it is obvious that in the “No-lens” case, the patterns in both E- and H-planes are not directional compared to all the cases when the lens is used. For the “Lens with ABS frame” case, it is evident that it deteriorates the radiation pattern, particularly in the H-plane, slightly increases the back-lobe radiation in both planes and also degrades the return-loss performance as shown in Figure 5.15(d). Additionally, it was observed during the fabrication process that the ABS plastic material was susceptible to deformation due to its high sensitivity to temperature. For the remaining two cases in which Rohacell® frames were used, aside from slight degradation of the S-parameters, it can be seen that the co-polar performance is not affected and is very similar to the reference case: “Lens with no frame”. This is probably due to the frame design and the lower ϵ_r values of the Rohacell® material (Rohacell® 300WF: $\epsilon_r = 1.3$ and Rohacell® 71HF: $\epsilon_r = 1.093$), compared to the ABS-plastic frame

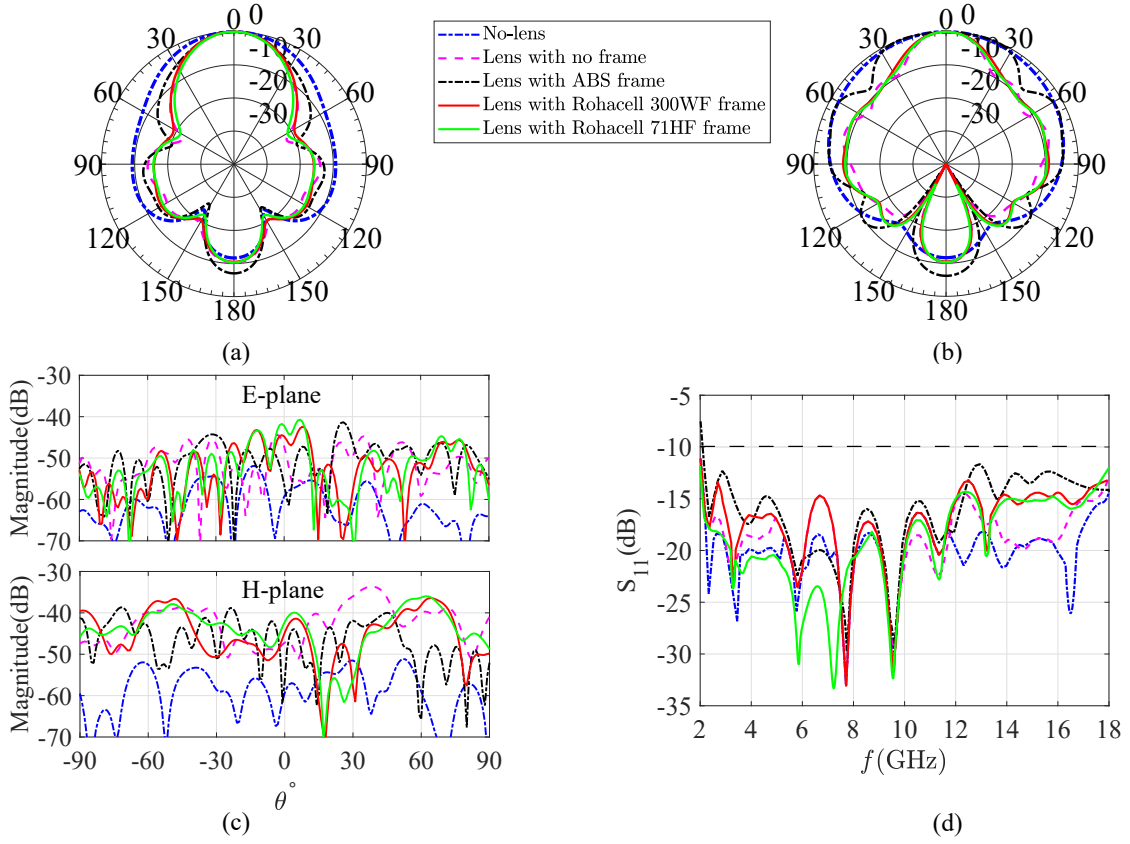


Figure 5.15: Effect of the lens and supporting frames on the radiation pattern at 3 GHz (a) E-plane Co-pol (b) H-plane Co-pol (c) E-plane X-pol (d) H-plane X-pol.

($\epsilon_r = 2.75$).

As depicted in Figures 5.15 (c) and (d), the cross-polarization patterns remain largely unaffected by any of these frames. Furthermore, it can be observed that when compared to the “No-lens case”, the use of the lens results in an increase in the cross-polarization levels in both the E- and H-planes. This phenomenon could be attributed to the minor gaps between the lens and the supporting frame due to the hexagonal shape of the outermost layer, and to the diffraction effects that occur at the edges of the lens, caused by slight over-illumination due to the phase center instability of the dual-ridged horn across the wide frequency band of operation. Furthermore, it could be due to the relatively higher dielectric value

of the outermost ring with $\varepsilon_r = 1.1$. Ideally, to avoid any reflections or diffraction effects at the edges, a foam material having $\varepsilon_r < 1.1$ is used for the outermost ring to have a smooth dielectric transition between the outermost ring and the free space. Based on these analyses, the Rohacell® frame was chosen compared to the ABS plastic frame. Among Rohacell® frames, Rohacell® 300WF was selected due to its lower cost, greater mechanical durability, and resilience to changes in temperature compared to the Rohacell® 71HF material.

5.4.3 Results and Validation

The finalized GRIN lens including the dual-ridged horn feed and the supporting frame was simulated using HFSSTM Integral Equation (HFSS-IE) solver that uses the MoM. The careful utilization of HFSS-IE with adaptive refinement meshing proves to be highly effective when simulation involves large structures. Based on the analysis discussed in the previous sections, the lens antenna was placed at a distance of $l_1 = FP = 177$ mm from the phase center (PC) of the source as shown in Figure 5.11(a). The radiation characteristics of the 3D-printed GRIN lens, along with the Rohacell® 300WF supporting frame, were assessed within the 3–18 GHz frequency band, in the far-field anechoic chamber as shown in Figure 5.14(e).

Figure 5.16 presents the sample simulated patterns in the principal planes for with and without the lens cases, at three specific frequencies. Figures 5.17 and 5.18 show the normalized measured co and cross-polar radiation patterns in the E- and H-planes, respectively, for both the lens and without the lens cases. For brevity, the comparison is shown for fewer frequencies. It can be seen that the proposed GRIN lens is effective in reducing the 3 dB beamwidth in both E and

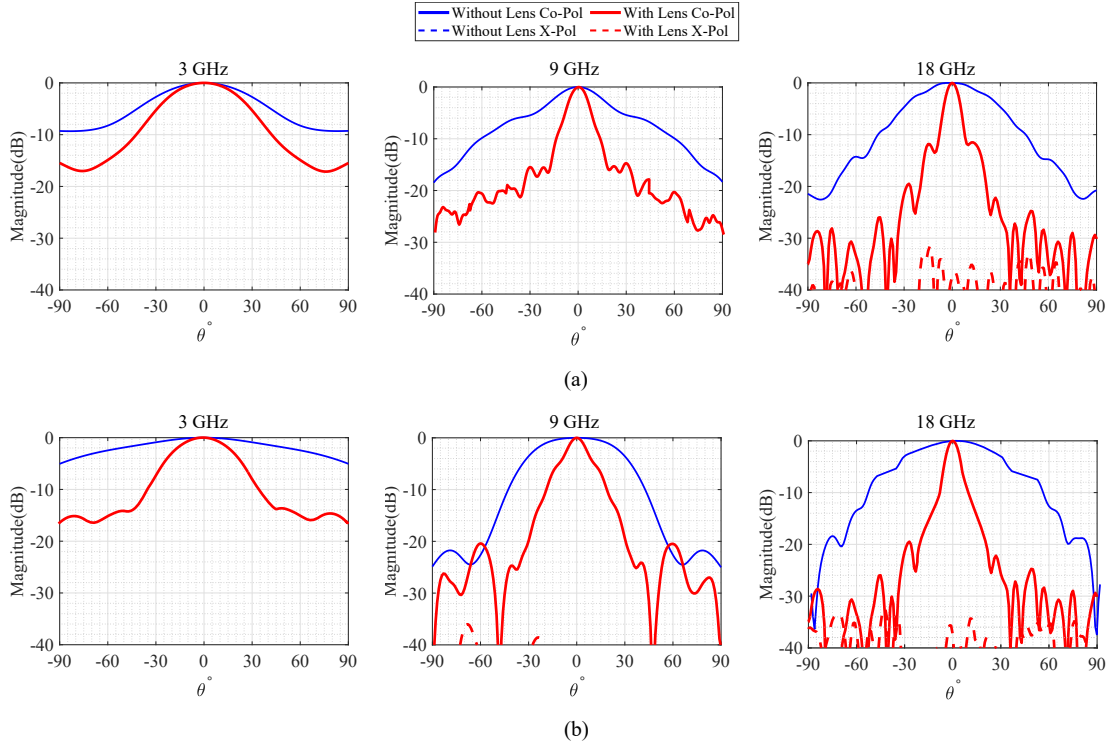


Figure 5.16: Sample simulated radiation patterns at several frequencies for with and without the GRIN lens cases (a) E-plane (b) H-plane.

H-planes, across the entire band of operation. However, in the case of the lens, there is a slight increase in the cross-polarization levels, particularly at higher frequencies ranging from 15 to 18 GHz. This phenomenon is common in both E- and H-planes. This can be due to several factors as discussed in the previous section. In order to demonstrate the performance through simulations, Overall a good agreement can be seen in the simulation and measured results in the entire 3–18 GHz band of operation.

Figure 5.19 shows the E-field (V/m) intensity plots in the cross-sectional E-plane for various frequencies. As anticipated, the proposed GRIN lens effectively converts the spherical wavefront generated by the dual-ridged horn into a plane wavefront, resulting in the collimation of the fields at the output. At higher of the

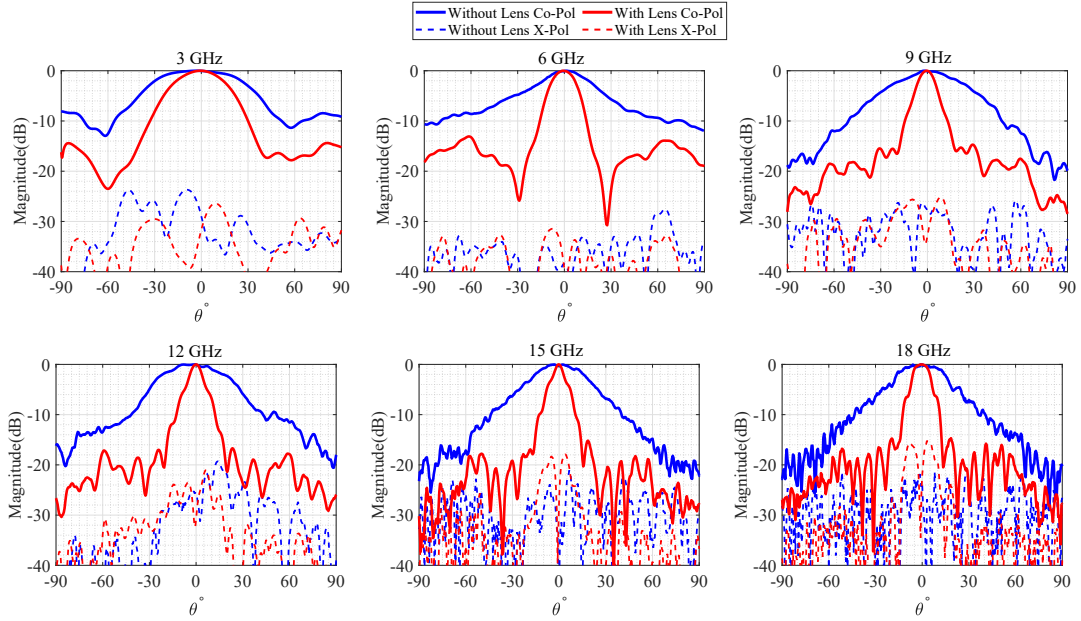


Figure 5.17: Measured E-plane radiation patterns at several frequencies for with and without the GRIN lens cases.

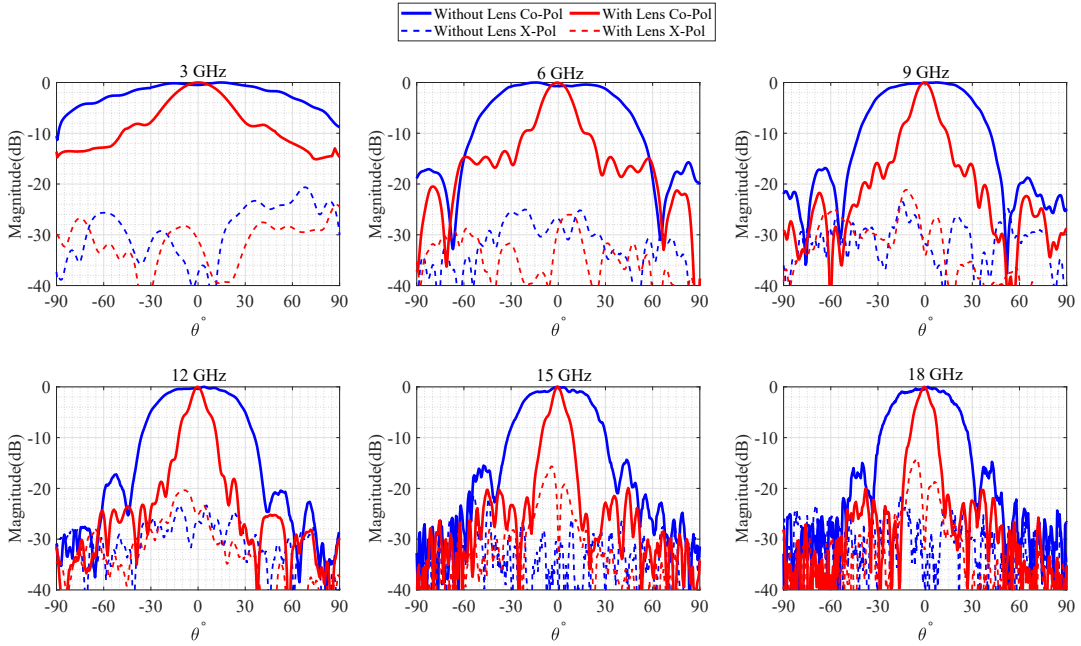


Figure 5.18: Measured H-plane radiation patterns at several frequencies for with and without the GRIN lens cases.

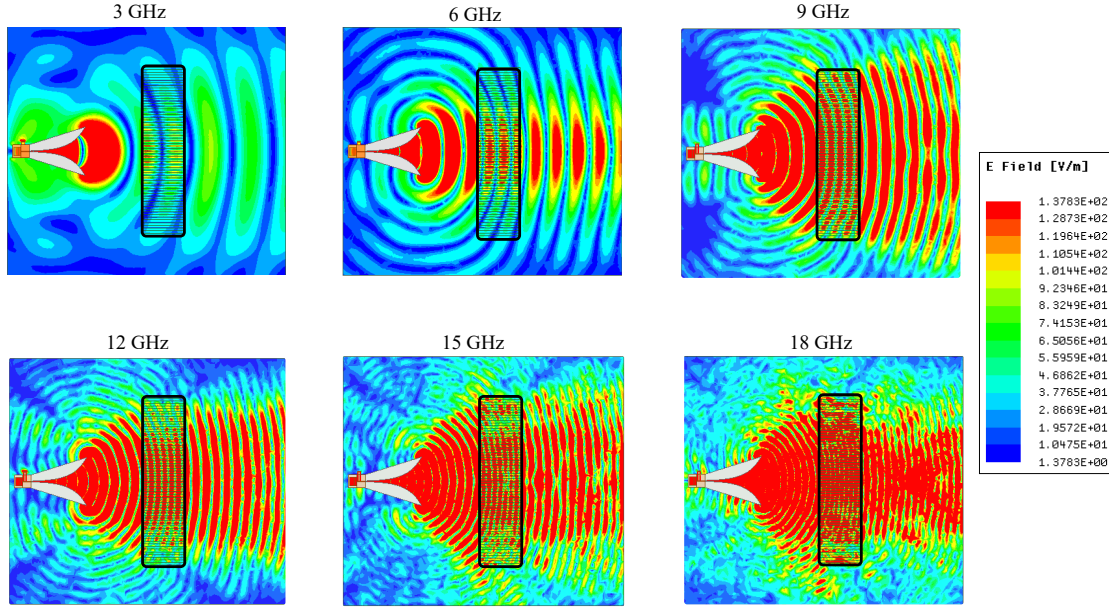


Figure 5.19: Cross-sectional view of the E-field (V/m) intensity plots in the E-plane for various frequencies.

band such as 15 and 18 GHz, it is evident that the influence of edge diffraction effects becomes more pronounced, thereby contributing to slightly higher levels of cross-polarization at these frequencies. In addition to the factors discussed earlier contributing to edge diffraction, these plots indicate that there is a little over-illumination of the lens occurring at these frequencies due to the dependence of the phase center of the feed antenna on the frequency. As reported in [110], the phase center of this dual-ridged horn antenna varies from 10 mm to 45 mm between 3–18 GHz, especially in the E-plane. To achieve proper illumination of the lens, the focal point of the lens must align with the phase center of the source antenna at the desired operating frequency.

Figure 5.20 compares the simulated and measured HPBW performance in both E- and H-planes for with and without the lens cases. It can be seen that without the integration of the proposed ADLA, the HPBW of the source antenna in the E and H-planes is around 75° and 120° , respectively, at the lowest frequency

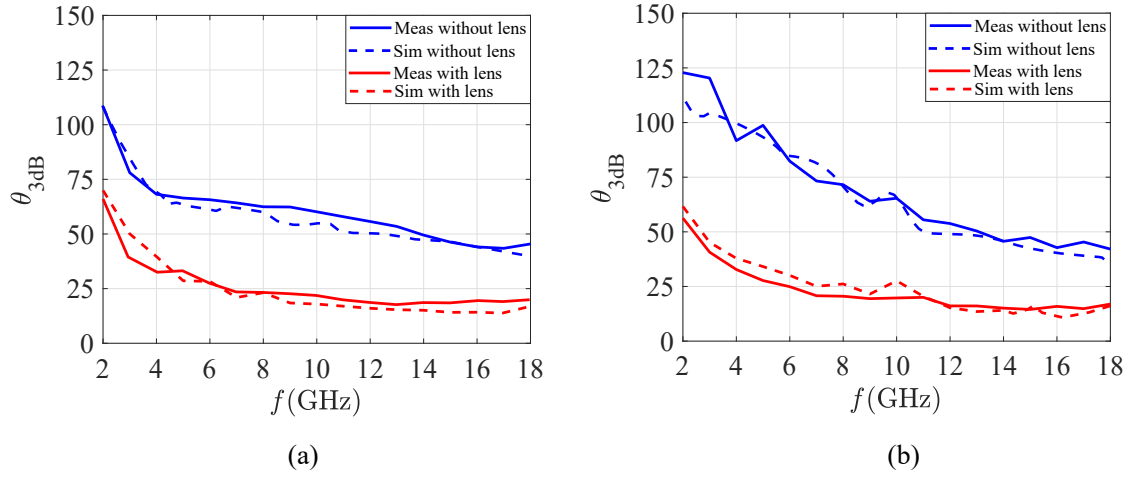


Figure 5.20: Measured and simulated 3 dB HPBW curves for with and without the GRIN lens cases (a) E-plane (b) H-plane.

of 3 GHz. Moreover, it can also be seen that there is significant variation in the beamwidth in both E- and H-planes across the band of operation. For instance, in the E-plane, the beamwidth varies from 75° to 40° , while in the H-plane, it varies from 120° to 40° , within the frequency band, representing a variation of 30.4% in the E-plane and 50% in the H-plane. Due to this discrepancy, the dual-ridged horn antenna itself fails to meet the specified beamwidth requirements ($\theta_{3dB} = 35^\circ \pm 5^\circ$) established for UAV-based radar characterization. In contrast, when the proposed GRIN lens is integrated, the HPBW is reduced to 35° at the lowest frequency of 3 GHz, and to 20° at 18 GHz, representing an average reduction of around 68% within the entire frequency band. Furthermore, with the proposed lens, the beamwidth varies from 35° to 20° , exhibiting a variation of around 27% across the frequency band. For both scenarios: with and without the GRIN lens, both the simulation and measured results exhibit an excellent level of agreement.

Figure 5.21 presents the HPBW patterns for both cases, capturing the symmetrical behavior in the E- and H-planes within a single plot. It can be seen that both simulation and measured results confirm the asymmetric radiation per-

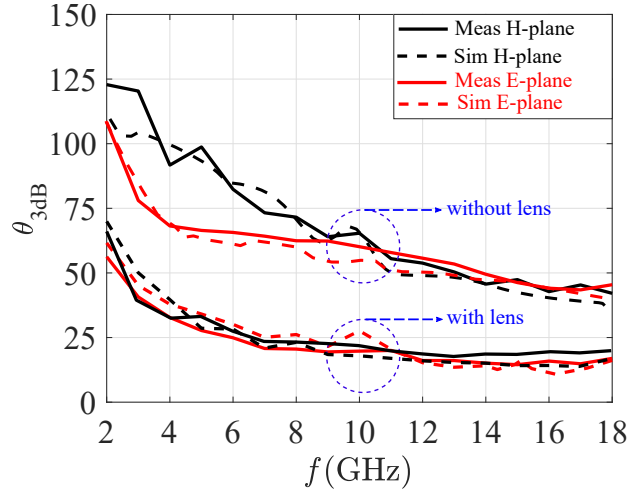


Figure 5.21: Comparison of symmetry in the measured and simulated E- and H-plane patterns for with and without the lens cases.

formance in the principal planes of the feed antenna when the lens is not used. However, upon integrating the proposed lens, the patterns demonstrate excellent symmetry in both the E- and H-planes. Both the simulation and measured results demonstrate the wide bandwidth and excellent beamwidth performance of the proposed lens, making it highly suitable for a diverse range of applications in radars, direction finding, and various communication systems.

Table 5.2 summarizes and compares the characteristics of the proposed design with other relevant works that focused on beamwidth reduction. It can be seen that the proposed work is the first to use a simple 3D-printed and low-cost lens solution for wideband beamwidth reduction compared to the complicated and costly corrugated or pinwall techniques. Additionally, the proposed work has the second highest absolute bandwidth after [153]. However [153] has used a complex technique of pinwalls, reported only H-plane beamwidth, and has a significant beamwidth variation of around 56%. Furthermore, it is important to highlight that, in contrast to other works, the proposed design with integrated lens encompasses the important S-band for weather radars and offers a consistent

Table 5.2: Performance comparison of the proposed technique with other works related to beamwidth reduction.

References	Frequency (GHz)			HPBW (Deg)		HPBW Variation (%)		HPBW Reduction Technique	Complexity	Cost
	f_{low}	f_{high}	Abs. BW	E-plane	H-plane	E-plane	H-plane			
[21]	2	12	10	15°-95°	15°-85°	± 73	± 70	Multi-layer Lens & Corrugation	High	High
[152]	4	16	12	20°-21°	18°-23°	± 2.4	± 12.2	Pin Walls	High	High
[153]	1	18	17	-	18.3°-65°	-	± 56	Modified Ridges & Pin Walls	High	High
[155]	4.5	18	13.5	29.5°-42.9°	28.2°-33.6°	± 18.5	± 10	Pin Walls	High	High
[156]	4.5	18	13.5	32.2°-42.6°	28.5°-33.5°	± 14	± 8	Pin Walls & Corrugation	High	High
[158]	7.5	18	10.5	40°-50°	42°-46°	± 11.1	± 4.5	Corrugation	High	High
[159]	8	18	10	22°-36°	30°-35°	± 24	± 7.6	Corrugation	High	High
SH2000-[110]	3	18	15	40°-75°	40°-120°	± 30.4	± 50	-	-	-
Proposed GRIN	3	18	15	20°-35°	20°-35°	± 27	± 27	3D-printed Lens	Low	Low

beamwidth of 35° in both E- and H-planes, thus satisfying the criteria for the UAV-based in-situ characterization of radars.

5.4.4 Proposed GRIN Lens: Summary

A UWB 3D-printed GRIN-ADLA is presented that enables the use of any wide-band antenna to be used on any UAV platform for in-situ characterization of radars and communication systems, which was not possible otherwise due to the wide beamwidth problem. The antenna is fully characterized and tested, and the simulation as well as measured results show promising results in terms of beamwidth reduction as well as pattern symmetry, especially at lower frequencies where achieving these characteristics, is challenging. It is shown that with the integration of the proposed GRIN-ADLA, radiation patterns in the principal E- and H planes are in excellent symmetry and the beamwidth is reduced from 120°-40° to 35°-20° within the desired 3-18 GHz frequency band, providing an average beamwidth reduction of 68%. The proposed technique of beamwidth reduction and achieving pattern symmetry is ultrawideband, effective, and low-cost compared to the narrow band and complex techniques presented in the literature.

In the coming sections, a few other lens and corrugated designs are presented to address the wide beamwidth problem of the QRHAs. However, these solutions are presented as a proof of concept and are only based on simulation results.

5.5 UWB Hyperbolic Lens

In this section, another solution to the wide beamwidth problem faced in ridged horn antennas is presented. The proposed UWB dielectric hyperbolic lens has a uniform ϵ_r . A QRHA serving as a feed for the lens is also designed having a size of $190 \times 160 \text{ mm}^2$. The proposed integrated system achieves narrow beamwidth in both E- and H-planes throughout the frequency of operation. For instance, at 3 GHz, the beamwidth is reduced by around 50% by integrating the proposed dielectric lens. The proposed integrated system has an absolute bandwidth of 20 GHz (1-20 GHz). Unlike the previously presented GRIN-ADLA, this hyperbolic lens provides ease of integration directly on the QRHA aperture thus providing more stability and compact overall size. The aperture opening of the QRHA is calculated such that the lens can be integrated exactly on the aperture of the QRHA with a focal point that coincides with the phase center of the QRHA. The proposed hyperbolic lens has a simple structure and can be fabricated using a commonly available Formlabs Form 3 resin-based 3D printer using a general-purpose grey resin (RS-F2-GPGR-04) having a ϵ_r of around 2.7).

Figure 5.22 shows the description of the proposed QRHA with an integrated dielectric hyperbolic lens antenna. The lens is symmetric in both planes thus helpful for its use with dual-polarized ridged horn antennas. The proposed lens is compact in size ($160 \times 160 \text{ mm}^2$) keeping in view targeting a lower frequency of 3 GHz. The lens is designed based on the theory presented in [173]. The size of

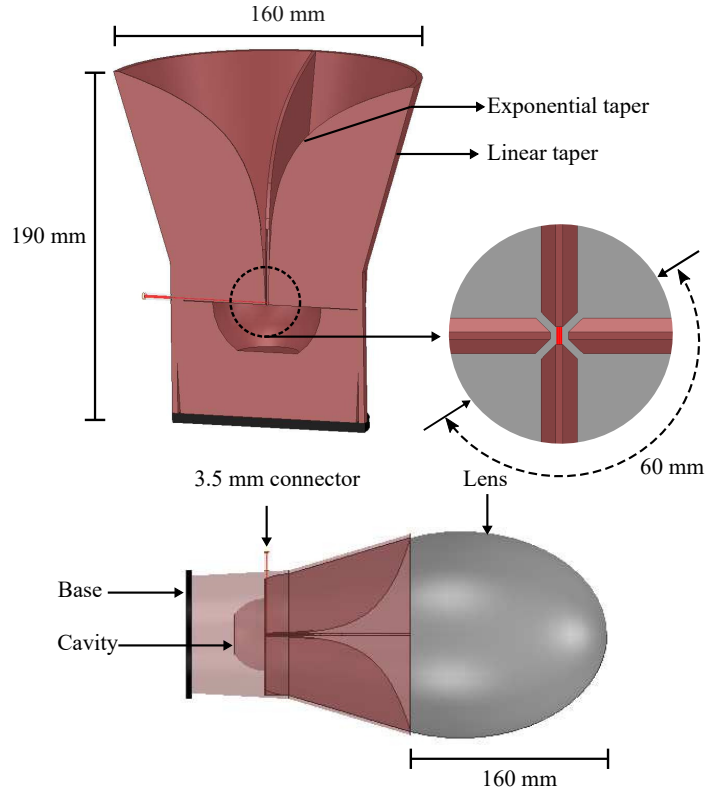


Figure 5.22: Geometry description of the proposed hyperbolic lens-integrated QRHA.

the proposed QRHA is $190 \times 160 \text{ mm}^2$.

Figure 5.23 shows the comparison of the HPBW of the proposed QRHA for with and without lens cases. It can be seen that reduction in beamwidth specifically at lower frequencies, is evident in both E and H-planes. At 3 GHz, the worst-case reduction in beamwidth is around 50%. From 2.5 GHz onward, the beamwidth in both planes is below the maximum required beamwidth for UAV-based applications.

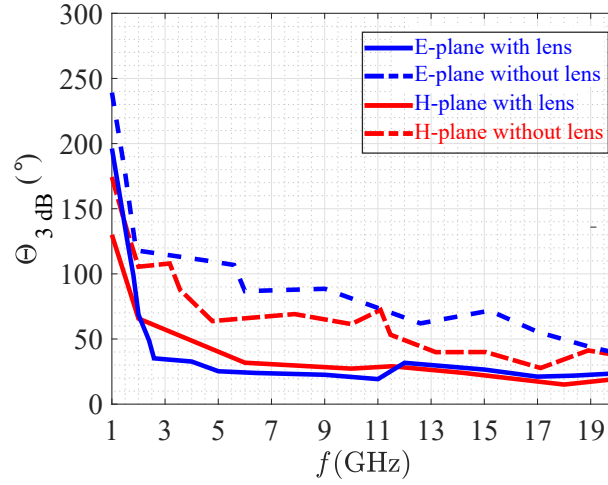


Figure 5.23: Simulated HPBW comparison in the principle planes for with and without the proposed hyperbolic lens structure.

5.6 Multi-Layer Lens and Corrugated Horn Antenna

In this section, two other beamwidth reduction techniques are discussed. The first technique involves the integration of a wideband multi-layer dielectric lens antenna (MDLA). In the second technique, corrugations are introduced in the DRHA to reduce the H-plane beamwidth for the entire band of operation. Both lens and corrugations reduce the H-plane HPBW by around 70% (for instance at 3 GHz) and show almost similar performance within the entire frequency band.

Figure 5.24 shows the geometry models of the standard (SH2000), proposed lens-integrated, and proposed corrugated dual ridged horn antennas. For the purpose of illustration, this figure also shows the E-field (V/m) distribution on the aperture of the antenna for each case, at 6 GHz. From the aperture field distributions, it can be observed that in the case of the SH2000 horn antenna shown in Figure 5.24(a), the fields are not symmetric in the E- and H-planes. If these fields are projected to a radiation pattern, it can be seen that the beamwidth is wider in the H-plane compared to the E-plane for a SH2000 horn antenna.

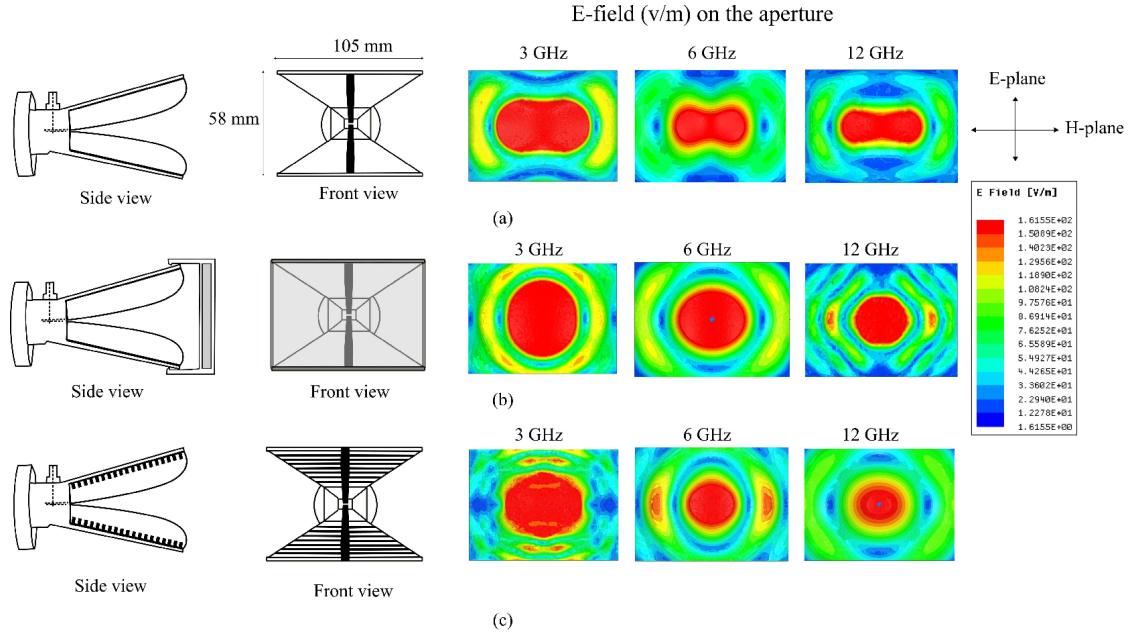


Figure 5.24: Antenna geometry models and E-field distribution on the antenna aperture at different frequencies (a) standard dual-ridged horn antenna without lens or corrugations (b) lens integrated dual-ridged horn antenna (c) corrugated dual-ridged horn antenna.

Figure 5.24(b) shows the case when the proposed MDLA is integrated with a dual-ridged horn antenna. The field distribution shows a nearly symmetric behavior in both E and H-planes. There is however some spillover radiation from the edges of the lens, which can be corrected by either slightly increasing the size of the lens or by slightly adjusting its focal point. For the third case, in which E-plane walls/plates of the standard dual ridged horn antenna are corrugated as shown in Figure 5.24(c), near-symmetric behavior is also observed compared to the SH2000 case.

Figure 5.25 shows the dummy lens-integrated ridged horn antenna prototype mounted on a DJI Matrice 600 drone. The profile of the multi-layer lens structure is also shown. The lens dielectric profile is symmetric where the central layer has the highest permittivity. This is inspired by a Luneburg-type lens configuration.

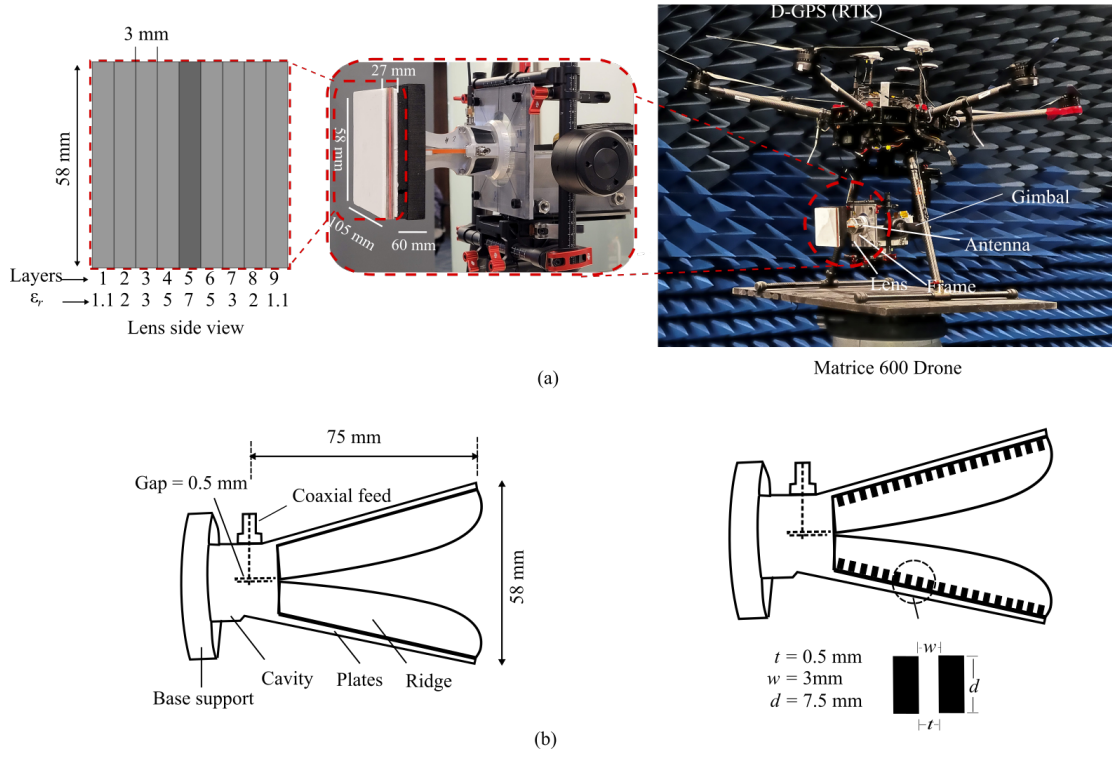


Figure 5.25: Illustration of the DJI Matrice 600 drone carrying the proposed antennas (a) a dummy prototype of the lens-integrated ridged horn antenna mounted on the drone (b) description and dimensions of the standard (SH2000) and proposed corrugated horn antennas.

The outer layers have permittivity values of close to the air so that the reflections between the air-dielectric layer are minimized. One important observation was made related to the dielectric profile of the lens. The smaller the dielectric gradient profile, the smaller are reflections between each layer, and thus high gain and low sidelobe levels can be achieved. However, this also depends on the availability of the material with the required values of the ϵ_r . To design such a material, an artificial technique can be used to achieve the desired values of the ϵ_r for each layer of the lens structure according to [19]. Figure 5.25 also shows a detailed description of the ridged horn antenna and the corrugation along with some of their dimensions. The antenna is fed using 50Ω coaxial feeding backed by the

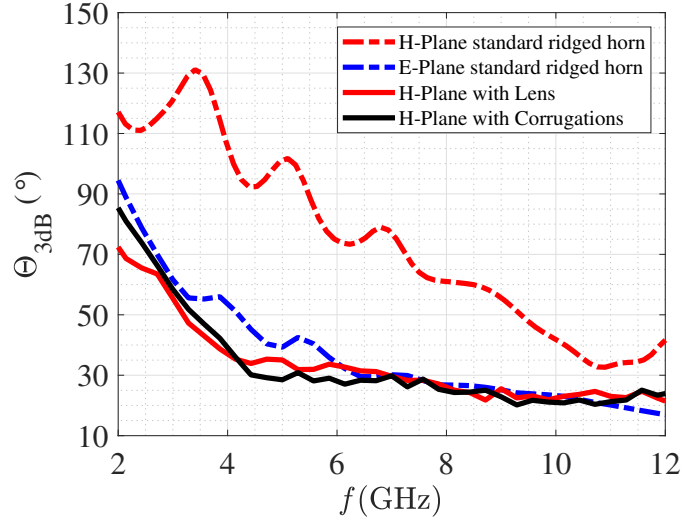


Figure 5.26: Comparison of the simulated 3 dB beamwidths for the three cases (for E-plane only one curve is shown since the lens or the corrugation does not significantly affect the E-plane beamwidth performance).

metallic cavity structure that improves the matching performance.

Figure 5.26 shows and compares the simulated 3 dB beamwidth curves of the H-plane for the three cases: standard SH2000 horn, ridged horn with an integrated lens, corrugated ridged horn. Since the E-plane performance is not significantly affected by any of these techniques, a single E-plane curve is shown to avoid redundancy. From the figure, it can be seen that both of these techniques reduce the H-plane beamwidth in the entire band of operation. For instance, the reduction is around 70% for the low-frequency band at 3 GHz. Overall, after a reduction in the H-plane beamwidth, the patterns seem to be nearly symmetric in the entire band of operation. This is also verified from the field distributions on the aperture as shown in Figure 5.24.

5.7 Summary

In this chapter, the results of a 3D-printed spherical lens were first discussed. several beamwidth reduction solutions were proposed to address the wide beamwidth challenge present in commercial or available ridged horn antennas. A 3D-printed GRIN-ADLA was then discussed in quite detail including its theory, design principles, simulation, as well as measurement results. The proposed GRIN-ADLA presented an interesting and feasible solution to the wide beamwidth problem of commercial ridged horn antennas for UAV-based characterization of radars. It was shown that with the integration of the proposed GRIN-ADLA, radiation patterns in the principal E- and H-planes were in excellent symmetry and the beamwidth was reduced from $120^{\circ}-40^{\circ}$ to $35^{\circ}-20^{\circ}$ within the desired 3–18 GHz frequency band, providing an average beamwidth reduction of 68%. Several other solutions such as hyperbolic and multi-layer stack-up lenses as well as corrugated ridged horn antennae were also presented. The hyperbolic lens showed a beamwidth reduction of around 50% at 3 GHz, while the multi-layer lens and the corrugated horn showed a reduction of around 70% at 3 GHz. All of these UWB beamwidth reduction solutions satisfied the beamwidth requirements set for in-situ characterization of radars using UAVs. In the following chapter, a standalone UWB QRHA is discussed that also satisfies UAV beamwidth requirements without the use of a dielectric lens structure.

Chapter 6

Ultrawideband Quad-Ridged Horn Antenna for Radars and Communication Systems

6.1 Introduction

Wideband antennas such as ridged horns, vivaldi, spiral, biconical, and log-periodic antennas have been widely used in EMC testing, satellite tracking, reflector feeds, radars, and other detection systems. Among these, horn antennas became a favorite choice especially in radars and EMC testing due to their high gain and directivity, relatively compact size compared to traditional horns, design simplicity, and simple feeding mechanism.

Horn antennas are rigid structures with straightforward feeding mechanisms and geometry construction, exhibiting high power handling capabilities in comparison to other antenna types. Nevertheless, the progress in wideband testing systems has limited the utilization of traditional horn antennas due to their limited bandwidth, highlighting the need for UWB horn antenna systems.

The introduction of dual- or quad-ridges inside the waveguide reduces the cutoff frequency when compared to a waveguide with identical width. These ridges, in turn, raise the cutoff frequencies of the subsequent two higher modes, enabling the creation of a waveguide capable of functioning across a frequency

range of 10:1 or greater [78]. If we employ this modified waveguide as the input for a horn and gradually taper the ridges until they no longer block the horn aperture, the resulting radiation pattern of the horn closely resembles that of a smooth-wall horn. Such horns are well-known as DRHAs having dual ridges or QRHAs consisting of four ridges inside the waveguide. Detailed theory and discussion on ridged horn antennas can be found in [2, 75–78].

6.2 Quad-ridged Horn Antenna: Challenges

Ridged horn antennas seem a promising antenna choice for a variety of UWB radar and communication system applications. However, in Chapter 3, it was discussed that due to the wide beamwidth problem of these ridged horn antennas, the radiation performance of the antenna probe is adversely affected in terms of the mismatch and cross-polarization, when the UAV platform is taken into consideration. These and some of the other challenges are discussed here. The main challenges are pattern distortion also known as pattern degradation at higher frequencies and wide beamwidth and poor impedance matching at lower frequencies [21–23, 25]. Moreover, the design of these antennas can get complicated if it is desired to simultaneously achieve lower values of VSWR < 2 , high cross-polarization levels of greater than -25 dB, narrow beamwidth performance especially in the lower region of the band, and having no pattern degradation.

6.2.1 Pattern Degradation

These antennas suffer from higher-order mode content which causes pattern distortion at high frequencies also known as pattern degradation. The extent of pattern degradation in this context is contingent upon the particular application.

For instance, if ridged horn antennas are used in near-field measurements, pattern degradation should not be a problem. However, in far-field measurements, EMC and communication system applications, radar characterization or calibration, and primary reflector feeds, where “well-matched” patterns are required, pattern degradation is not acceptable. This significant degradation, typically more than 4 to 5 dB in patterns, renders these antennas less suitable for the aforementioned applications.

A solution to this problem is a compromise between the maximum achievable bandwidth and pattern degradation. Commercially available ridged horn antennas adopt smaller cavity opening diameters to achieve maximum single-mode operation by shifting the higher mode TE_{31} away from the desired frequency band. However, this approach has a trade-off, as it also shifts up the propagation of the first dominant mode TE_{10} , resulting in poor matching characteristics at the lower frequency band. Vice versa, if the cavity opening diameter is increased to have good matching in the lower band, it also shifts down the TE_{31} mode, and thus the pattern degradation is seen in the desired band. This is one of the biggest challenges associated with QRHAs. This is explained in Figures 6.1, 6.2, and 6.3. It should be noted that the mode TE_{21} is not shown since it is canceled due to the feeding symmetry [174]. *Due to these challenges, the majority of the literature on ridged-horn antennas focuses on frequencies starting from 6 or 8 GHz and beyond. Designs targeting lower frequencies of 1 or 2 GHz often experience either poor matching or significant pattern degradation, particularly at higher frequencies. Because of this, the commercial QH2000 antenna uses a small cavity diameter to remove pattern degradation at the expense of poor VSWR performance (VSWR >3-4) at lower frequencies from 1-4 GHz.*

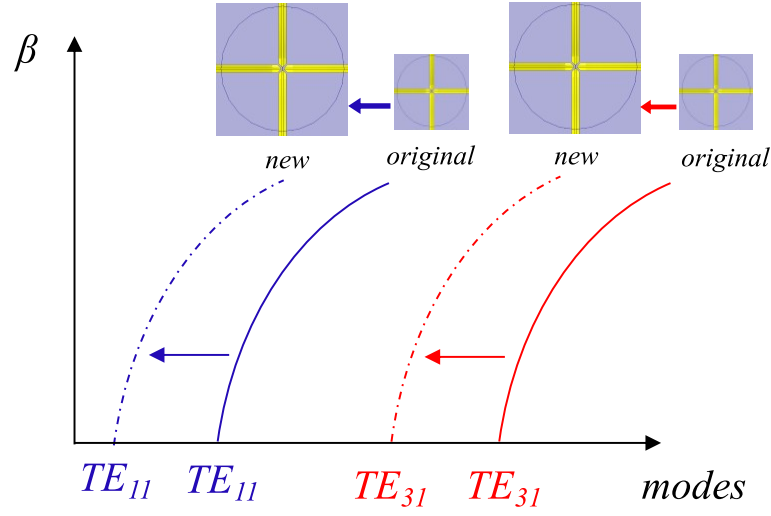


Figure 6.1: A conceptual diagram showing the effect of the cavity opening diameter on the modes propagation on a fictional frequency scale.

6.2.2 Wide Beamwidth Performance

The beamwidth of a horn antenna, or any aperture antenna in general, exhibits an inverse relationship with the size of the aperture in terms of wavelength. QRHAs typically suffer from wide beamwidth and beamwidth instability problems, especially in their lower region of the operational band i.e. between 1-6 GHz. As an example, the commercially available QRHA in [18] has wide beamwidth values of 120° – 80° corresponding to the frequency range of 3-5 GHz, where 120° corresponds to 3 GHz and 80° corresponds to 5 GHz. As shown in Chapter 3, such wide beamwidth performance of these QRHAs limits their use in some important applications such as in UAV-based in-situ characterization of radars. [22, 23].

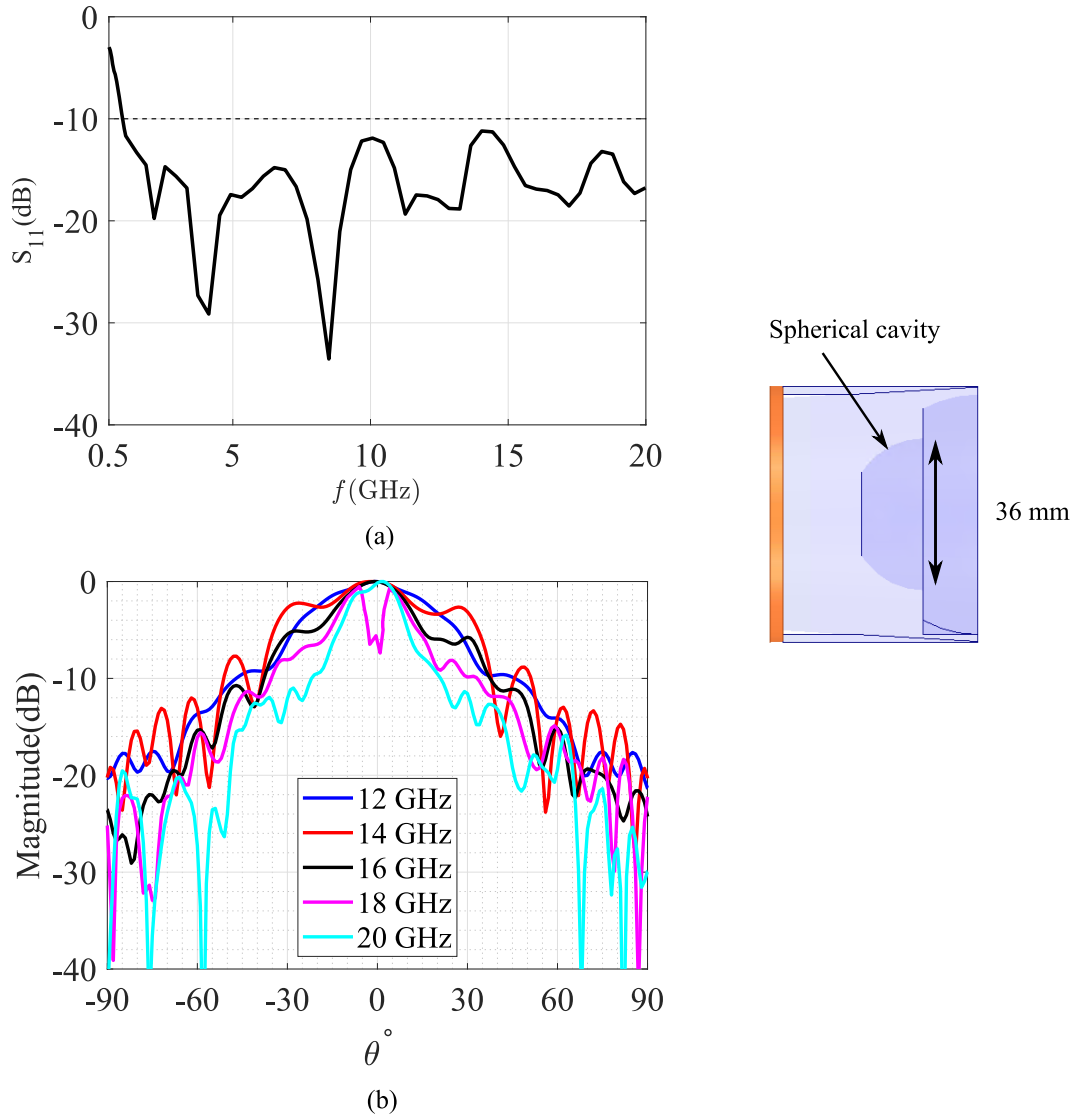


Figure 6.2: Effect of the cavity opening diameter on the S-parameter and pattern degradation performance (a) S-parameter curve showing the return loss is greater than 10 dB at 1 GHz (b) 2D radiation pattern with pattern degradation at 18 GHz.

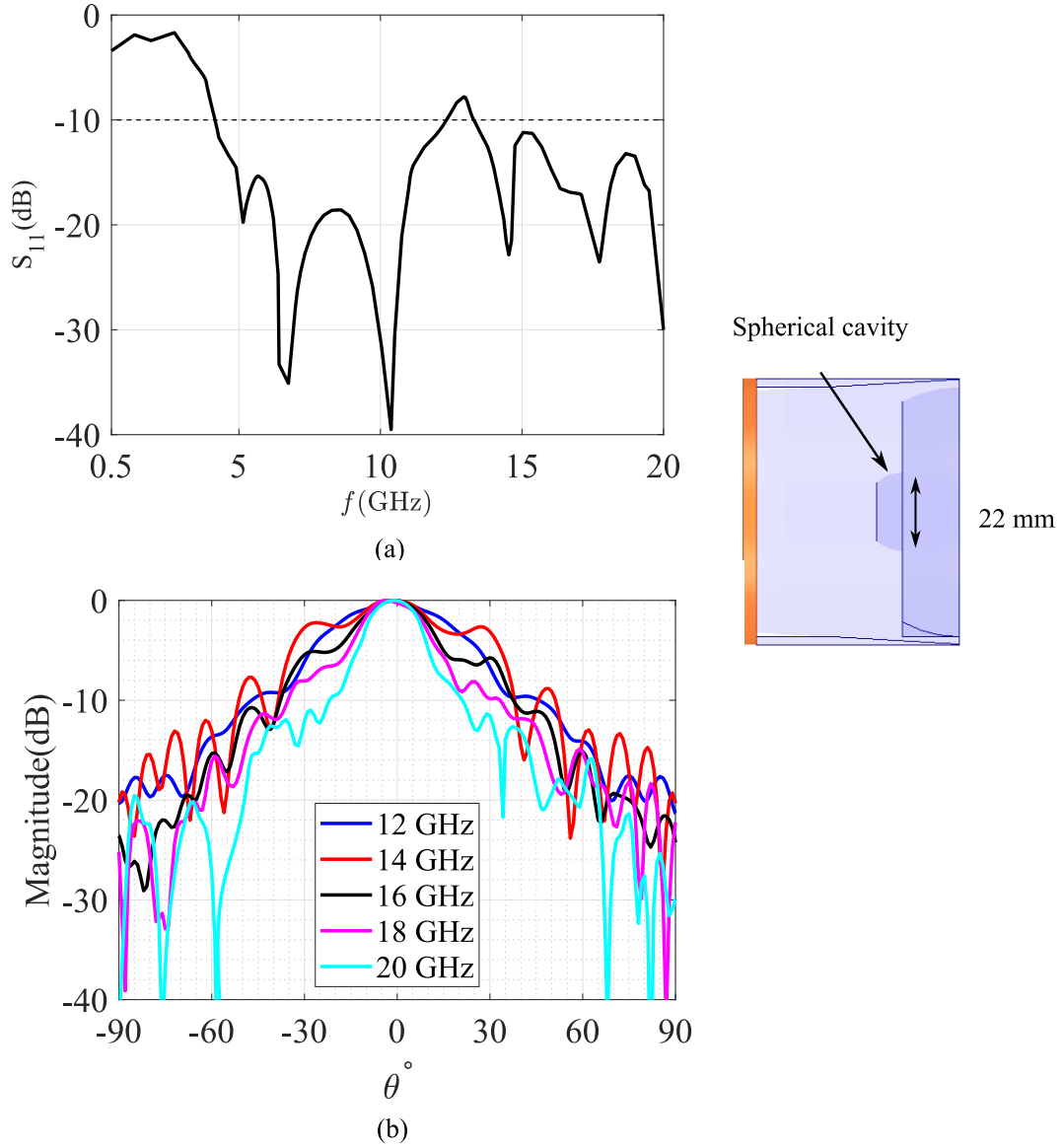


Figure 6.3: Effect of the cavity opening diameter on the s-parameter and pattern degradation performance (a) S-parameter curve showing the return loss is greater than 10 dB at 4 GHz (b) 2D radiation pattern with no degradation at 18 GHz.

6.3 Design Guidelines

6.3.1 Selection of the Ridge Profile

The ridge profile is one of the important parameters that significantly affects the VSWR performance. The most common ridge profile used in literature is based

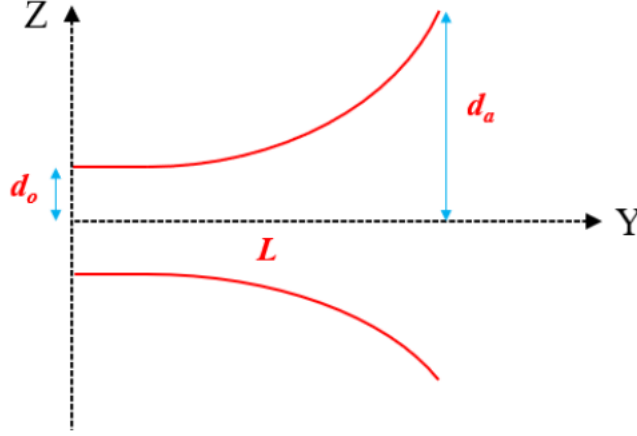


Figure 6.4: Demonstration of the ridge profile using original exponential equation.

on an exponential equation which is expressed as [175]:

$$Z(Y) = d_o \times e^{kY} \quad (6.1)$$

in which $0 \leq Y \leq L$, and k is expressed as:

$$k = \frac{1}{L} \ln \frac{d_a}{d_o} \quad (6.2)$$

where L represents the length of the horn section, d_o is the spacing or origin offset, and d_a is the radius of the horn aperture. For the purpose of illustration, these parameters are shown in Figure 6.5. It is well-known that the size of the aperture of the horn dictates the beamwidth performance at the lowest frequency of the band. In order to increase the size of the aperture to achieve beamwidth reduction, ridge profile based on equation (6.1) does not serve the intended purpose. For instance, if d_a is increased, the impedance of the waveguide transforms abruptly

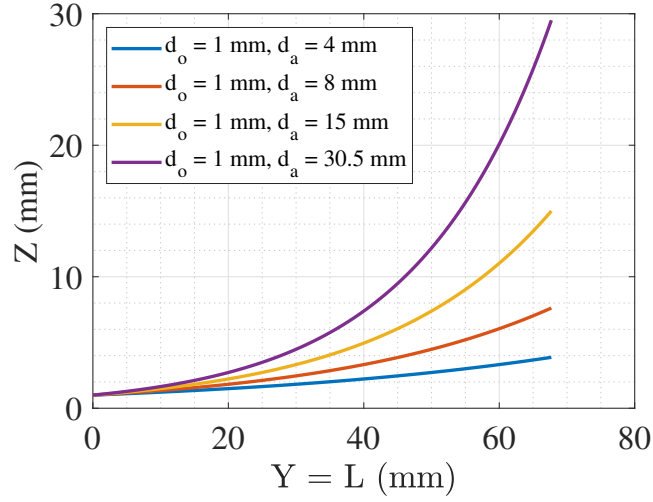


Figure 6.5: Ridge profiles using the original exponential equation with various aperture sizes.

to the free-space impedance which degrades the VSWR performance [175]. This phenomenon is shown in Figure 6.5. To achieve a smooth impedance transition along with increased aperture size, there is a need for a slight modification in the ridge profile equation. The complete ridge profile is divided into two segments as shown in Figure 6.6. The modified profile can be represented by the following set of expressions:

$$Z_1(Y) = d_o \times e^{k_1 Y} \quad 0 < Y < L_1 \quad (6.3)$$

$$Z_2(Y) = d_t \times e^{k_2 Y} \quad L_1 < Y < L_2 \quad (6.4)$$

where k_1 and k_2 can be expressed as:

$$k_1 = \frac{1}{L_1} \ln \frac{d_t}{d_o} \quad (6.5)$$

$$k_2 = \frac{1}{L} \ln \frac{d_a}{d_t} \quad (6.6)$$

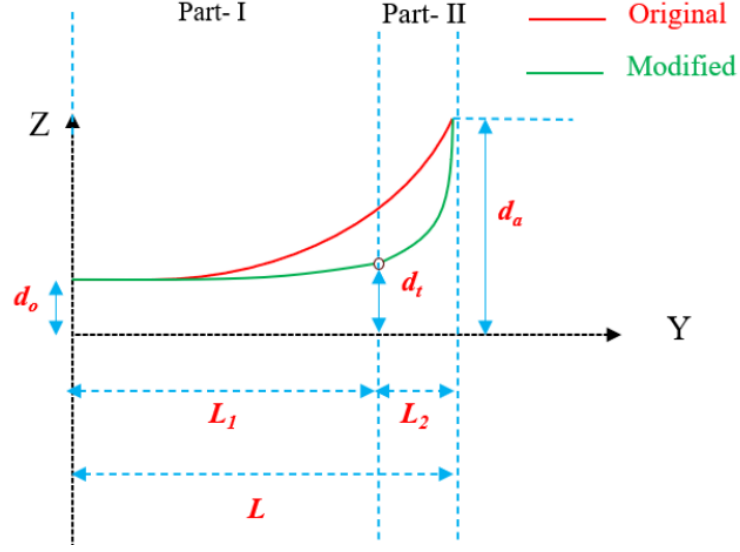


Figure 6.6: Demonstration of the ridge profile using modified set of equations.

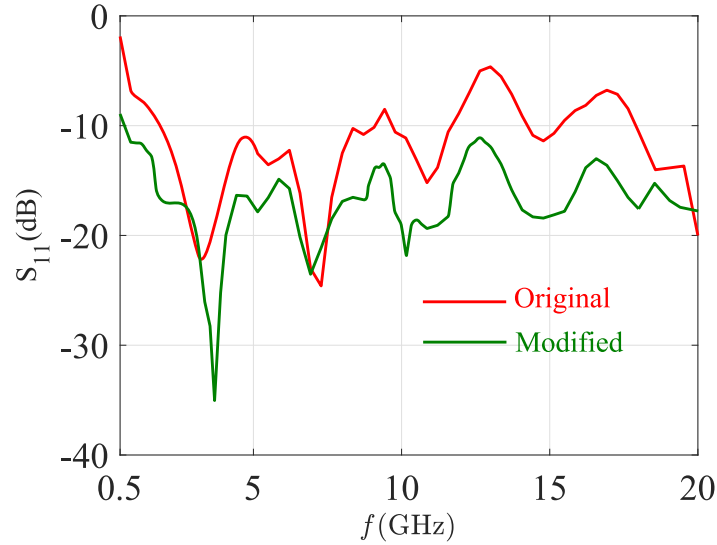


Figure 6.7: Illustration of the S-parameter performance comparison for ridge profiles using original and modified equations.

Figure 6.7 shows the comparison of the S-parameters curves for ridge profiles using the original exponential equation and the modified equation. It can be seen the ridge profile has a significant effect on the matching performance. QRHA using a modified ridge profile has a better matching compared to the one using

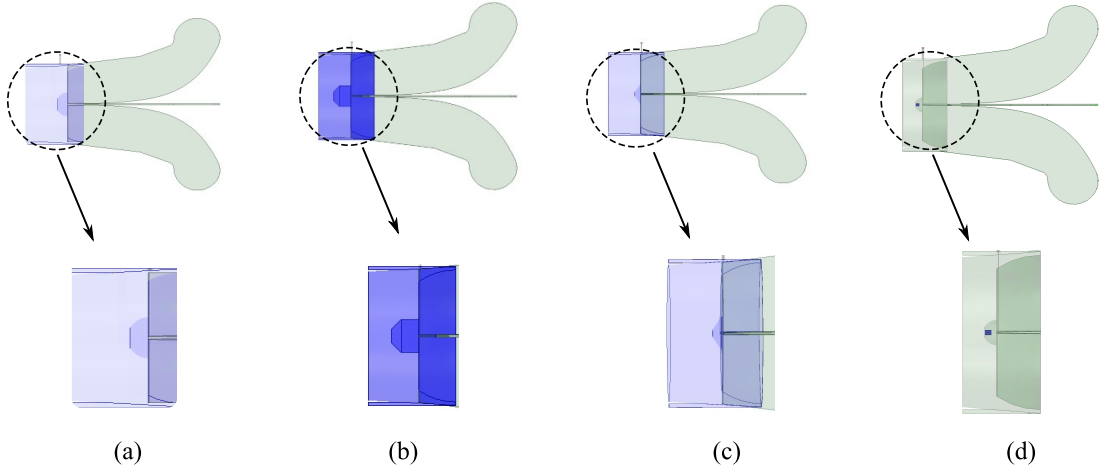


Figure 6.8: QRHA with different cavities in the feeding section (a) spherical cavity (b) pyramidal cavity (c) conical cavity (d) spherical cavity with a metallic post.

the original equation.

6.3.2 Selection of the Cavity

Several simulations were carried out to determine the right cavity for the proposed QRHA. Figure 6.8 displays QRHA configurations utilizing different types of cavities, including spherical, pyramidal, conical, and spherical cavity loaded with a metallic post. Figure 6.9 shows the S-parameter performance for these cases. It can be noticed that the conical cavity shows comparatively better matching performance when compared to other cases. Moreover, the conical cavity has a lower cut-off frequency of 0.7 GHz at -10 dB, compared to other cases. For these reasons, the conical cavity is selected for the final QRHA design to be discussed in the coming section. It is important to note that the cavity type has minimal impact on the radiation performance. Consequently, for the sake of brevity, a comparison of the radiation patterns for these different cases is not shown here.

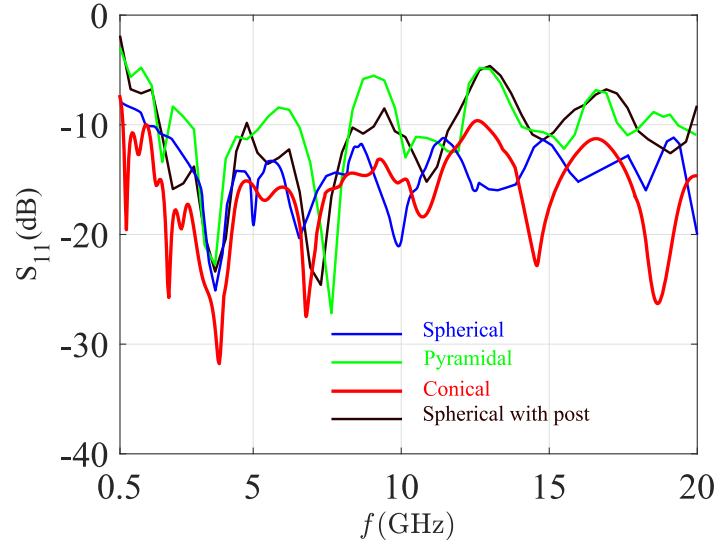


Figure 6.9: Comparison of the S-parameter performance for different cavities.

6.3.3 Selection of the Aperture Size

As discussed previously that the size of the aperture is related to the beamwidth performance. Based on the targeted application, the size of the aperture can be optimized to achieve the desired HPBW performance. Figure 6.10 shows the comparison between two QRHAs with varying aperture sizes. It is evident that the larger aperture size results in narrow beamwidth performance across the entire operational frequency band. For instance, it can be seen that at 3 GHz, when $D = 200$ mm, the HPBW is 70° and 74° , in the E- and H-plane, respectively. When $D = 276$ mm, the HPBW is reduced to 52° and 54° , in the E- and H-plane, respectively.

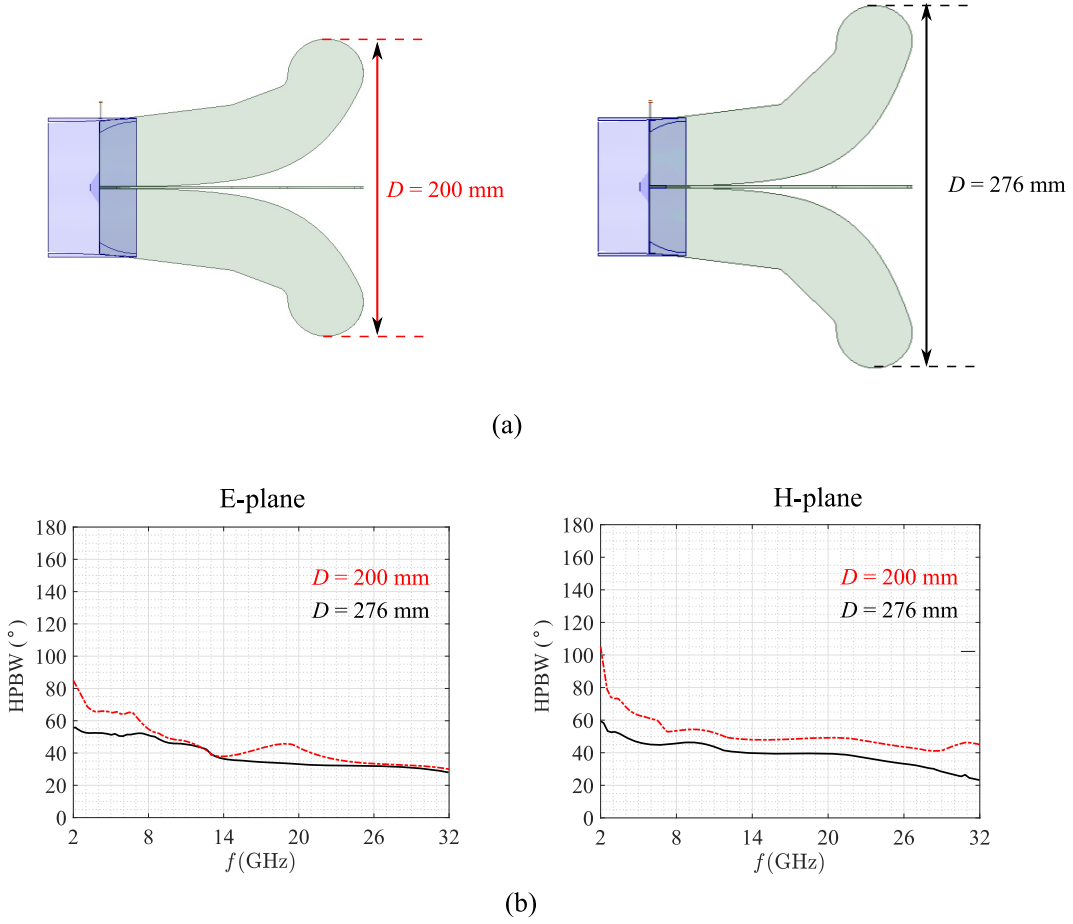


Figure 6.10: Effect of the aperture size on the beamwidth performance (a) QRHAs with two different aperture sizes (b) HPBW comparison in the E- and H-planes.

6.4 Results and Discussion

Figure 6.11 shows the geometry description of the proposed open boundary QRHA having a total size of $(296 \times 276 \text{ mm}^2)$ including the supporting frame. The antenna dimensions are kept according to its operation at the lowest frequency of 1 GHz. The overall geometry consists of three main sections: quad-ridges, cavity section, and supporting frame as shown in Figure 6.11. The all-metal design can be manufactured using standard CNC milling technique. The cavity diameter, ridge profile, cavity flaring, and the integration of the ridges at the cavity are optimized

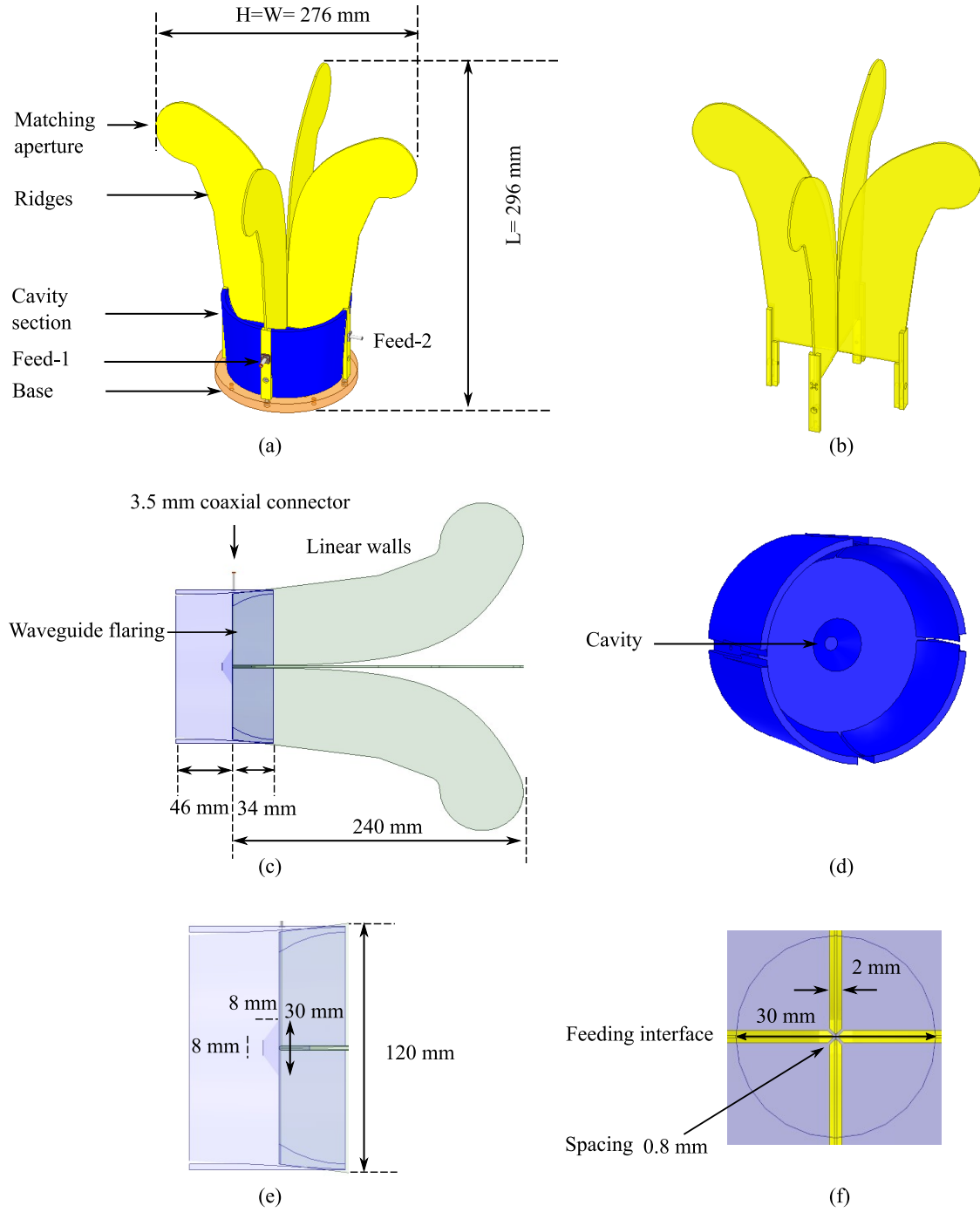


Figure 6.11: Geometry description of the proposed QRHA (a) isometric view of the complete geometry (b) isometric view of the quad-ridges (c) side view of the proposed QRHA with dimensions (d) cavity (e) side view and dimensions of the cavity (f) feeding interface at the cavity .

to have a maximum achievable bandwidth as well as no pattern degradation. The aperture size ($D = 276$ mm) is chosen based on the analysis discussed in the previous section.

The antenna geometry is modeled and simulated in HFSSTM using a high-performance computing (HPC) domain memory solver to reduce the simulation time. All the parameters of the design such as the cavity opening diameter, cavity bottom diameter, cavity depth or height, waveguide flaring, feeding or ridge spacing, feeding location, ridge thickness, and horn length are optimized to achieve the maximum possible bandwidth as well as good return loss performance. Figure 6.12 shows the return loss performance of the proposed QRHA for one port excitation. It can be seen that the antenna has a return loss of more than 10 dB for the entire frequency band of operation and excellent ultrawideband performance between 1-32 GHz. Figure 6.13 shows the VSWR performance of the proposed QRHA. It can be noticed that the proposed QRHA has a desired VSWR performance ($VSWR < 1.92$) in the interested frequency band between 1-32 GHz.

Figure 6.14 shows the simulated two-dimensional normalized radiation patterns of the proposed QRHA for sample frequencies within the entire frequency band. The patterns are shown for both E- and H-planes as well as for co- and cross-polarizations. For brevity, results for only one port are shown here. It can be seen that the proposed QRHA has a very good radiation performance in both lower and upper-frequency bands. Notably, this ridged-horn antenna demonstrates no pattern degradation even at higher frequencies due to the careful optimization of the cavity and the feeding section. This pattern degradation is a common concern for most of the ridged-horn antennas designed to operate at the lowest frequencies within the 1-6 GHz range while aiming for well-impedance matching performance

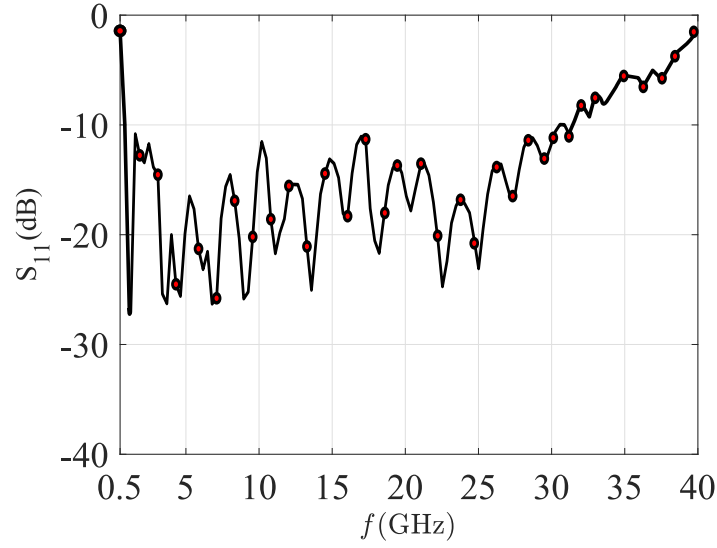


Figure 6.12: S-parameter performance of the proposed QRHA (for simplicity results of only one port are shown).

below -10 dB at those frequencies. It can also be noticed that the proposed QRHA has very low cross-polarization levels of greater than -30 dB in the 1-18 GHz band and a fair cross-polarization of around -20 dB in the 20-32 GHz region of the band.

The proposed design carefully employs the maximum achievable cavity opening diameter of 30 mm before pattern degradation becomes noticeable. This gives a maximum possible bandwidth spanning from 1-32 GHz and excellent matching below -10 dB without having pattern degradation at higher frequencies. Due to these features, the proposed design holds great potential for a diverse range of applications, in EMC testing, radar calibration, and other communication systems operating anywhere between 1-32 GHz.

Figure 6.15 shows the HPBW performance comparison of the commercial QRHA and the proposed antenna in both E- and H-planes. Significant reduction in the beamwidth can be seen in both planes especially in the lower region of the frequency band between 2-8 GHz. It can be noticed that the commercial

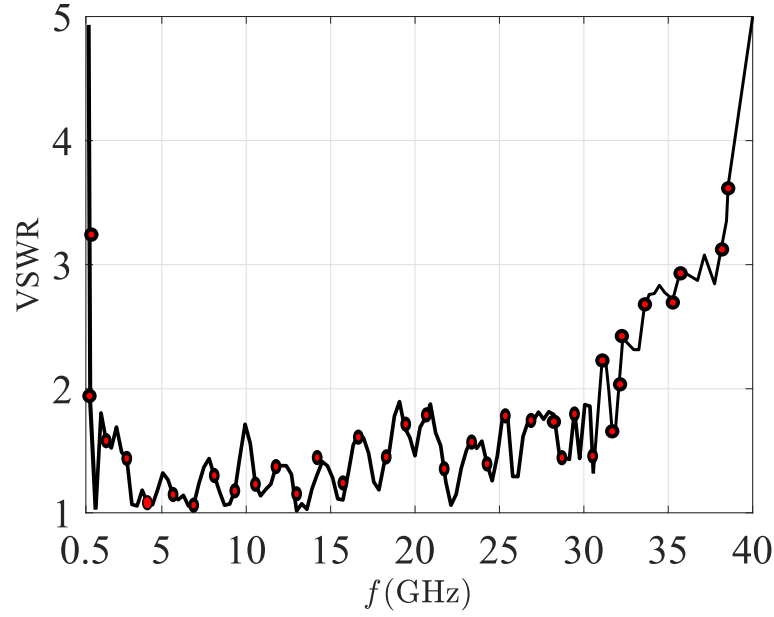


Figure 6.13: VSWR performance of the proposed QRHA.

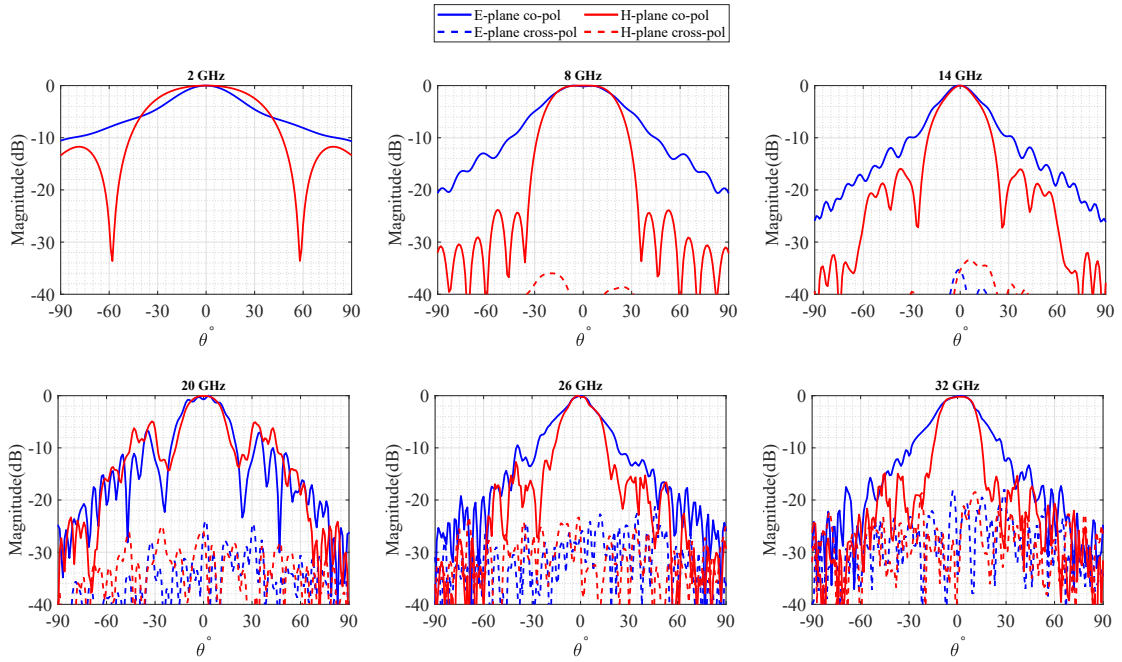


Figure 6.14: 2D normalized E- and H-plane radiation plots of the proposed QRHA at various frequencies within the operational band (for the sake of brevity, patterns from only one port are displayed).

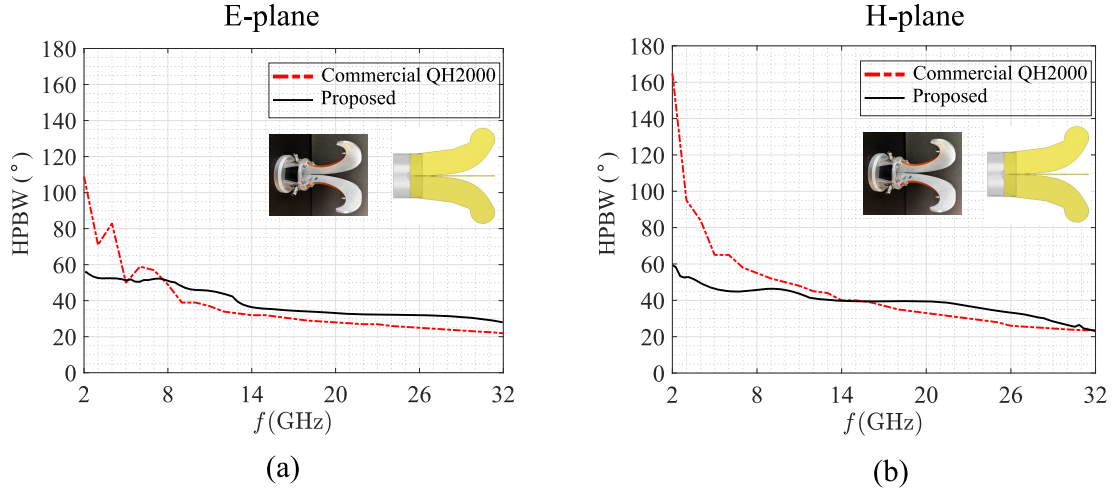


Figure 6.15: HPBW comparison of the commercial QH2000 and the proposed QRHA (a) E-plane (b) H-plane.

Table 6.1: Performance comparison of the commercial QH2000 and the proposed QRHA-1.

	Freq. (GHz)	HPBW (°)		RL (dB)	Size (mm)
Probe Type	(f_{min}/f_{max})	E-plane (f_{min}/f_{max})	H-plane (f_{min}/f_{max})		(H-W-L)
QH2000	2/32	110/23	165/23	> 6	110 x 105 x 105
Proposed	2/32	58/30	60/23	> 10	276 x 276 x 296

QH2000 [18] antenna has a beamwidth of $110^\circ/23^\circ$ at 2/32 GHz in the E-plane, while $165^\circ/23.5^\circ$ at 2/32 GHz in the H-plane. The proposed QRHA-1 has a beamwidth of $58^\circ/30^\circ$ at 2/32 GHz in the E-plane and $60^\circ/23^\circ$ at 2/32 GHz in the H-plane. This shows a reduction of 47% and 64% at 2 GHz in E- and H-planes, respectively. Table 6.1 summarizes the performance of both of these antennas. It should be noted that the size of the proposed antenna is larger due to the beamwidth requirements in the lower region of the band. In general, the suggested design surpasses the commercial design in terms of -10 dB impedance bandwidth, pattern degradation, and beamwidth performance.

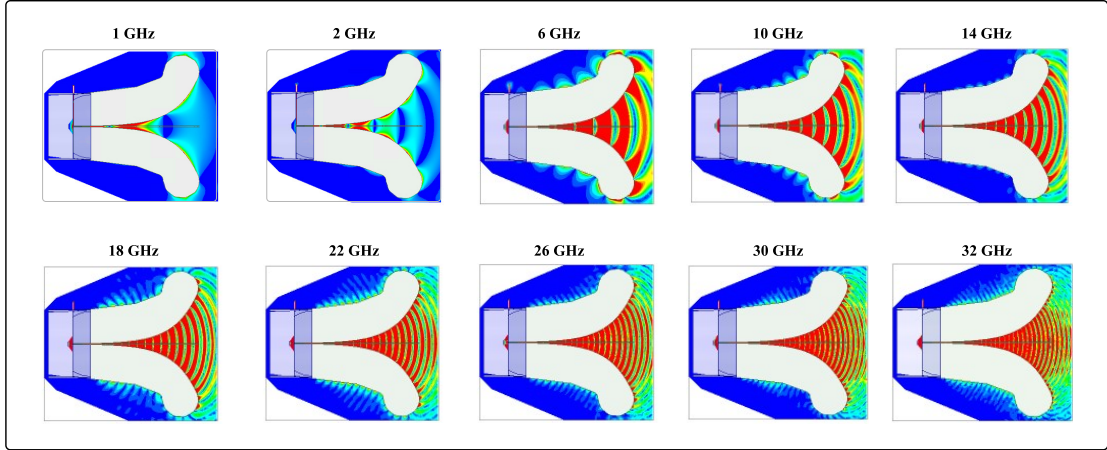


Figure 6.16: E-field (V/m) distribution in the cross-sectional plane of the proposed QRHA at various frequencies of the band.

Figure 6.16 shows the cross-sectional view of the magnitude of the E-field (V/m) distribution at various frequencies. It can be seen that at low frequencies, the field is loosely confined between the ridges, especially at the aperture of the QRHA. However, at higher frequencies, the field is well confined between the ridges. This also shows that at higher frequencies, the majority of the field's concentration is in the central region of the QRHA.

The power handling capability of the proposed all-metal QRHA is dictated by the maximum input power of the connector. As shown in Figure 6.11, the proposed design uses a 3.5 mm precision connector. This connector incorporates an air dielectric, resulting in superior performance compared to the Teflon-based connectors. While these 3.5 mm connectors can operate effectively up to 38 GHz, they are commonly rated by some companies up to 34 GHz or 26.5 GHz [176]. It is important to remember that specific application will determine the frequency range and power handling requirements of the connector. The connector's power handling is dependent on factors such as its construction, temperature, and reflected power. A connector's continuous CW power rating is mainly based on

the temperature rise, due to power dissipation resulting from ohmic and dielectric losses. The maximum input power for the 3.5 mm connector at its maximum frequency of 34 GHz, is 45 Watts [176, 177].

6.5 Summary

Although the ultrawideband performance of QRHA makes it a suitable candidate for EMC testing and radar calibration, a compromise is usually made between the radiation performance (pattern degradation) at more than 12 GHz and the matching performance in the lower band from 1-6 GHz. In this chapter, challenges associated with existing QRHAs were discussed in detail. A new design was proposed that overcame these limitations by carefully optimizing the cavity section, cavity flare section, ridge flaring profile, and along with other feeding parameters. The proposed QRHA showed exceptional bandwidth performance, ranging from 1 to 32 GHz, reaching as low as 1 GHz, and maintaining an impressive return loss of more than 10 dB across the entire frequency band. Moreover, it exhibited favorable cross-polarization performance and no pattern degradation even at higher frequencies within the band. These qualities highlight its potential for diverse applications, including EMC testing and radar calibration. The following chapter summarizes the main contributions and research outcomes of this dissertation followed by possible future research directions.

Chapter 7

Conclusions

7.1 Summary

The increasing demand for UWB antenna systems is driven by the critical role they play in the success of future multi-functional communication, radar, and sensing systems. These antennas have a wide range of applications, including EMC testing, medical imaging, through-wall imaging, satellite tracking, and weather radar calibration, among others. This dissertation is tailored to address the design of UWB antenna systems for UAV-based radar calibration, aiming to address the current challenges related to bandwidth and radiation performance. It encompasses the theory, design, and modeling of a UWB “one-size-fits-all” standalone or integrated antenna system satisfying the beamwidth, symmetry, as well as bandwidth requirements set for the calibration of weather radars.

Previous studies related to UAV-based in-situ calibration of weather radars have used very narrow band antenna probes having a bandwidth of only 100 MHz. Since each antenna behaves differently with the gimbal and the UAV platform, the use of these narrowband antenna probes requires several dedicated UAV-based calibration systems for radars operating in the L-, S-, C-, X-, Ku-, and K-bands, thus heavily contributing to the overall cost of the system. Moreover, this also

entails using and subsequently re-calibrating multiple probes, leading to time and greater labor involvement. Consequently, there is a strong desire for a single UWB antenna thus a single UAV calibration system that, once calibrated with the drone, can be employed for radar characterization across multiple frequency bands, thus saving time, cost, and labor efforts.

For the purpose of finding an ideal antenna probe that simultaneously provides UWB performance along with satisfying the pattern mismatch, pattern symmetry, as well as beamwidth requirements, the commercially available UWB QRHA: QH2000 is rigorously tested for its RF performance. The detailed analysis and measurement results are reported in Chapter 3. It is shown that although these antennas are lightweight and have UWB performance, they do not satisfy the probe antenna requirements set of UAV-based calibration of weather radars. These commercially available QRHAs exhibit a very wide beamwidth performance, typically ranging from 150° – 80° in the lower part of the frequency band from 1-6 GHz. This wide beamwidth characteristic limits their suitability for a critical portion of the frequency spectrum for instance at 3 GHz, which is primarily employed by weather radars and communication systems. It is shown through measurements that the extensive beamwidth radiation performance leads to unwanted scattering from the UAV structure, resulting in distortions in the co-polar patterns and impacting the cross-polarization performance.

Chapter 4 establishes the groundwork for the design of artificial dielectric materials and lenses. It presents the theory, design, manufacturing, and testing of artificial dielectric material tailored for low $\tan(\delta)$ and ε_{eff} values suitable for microwave and mm-Wave applications. In this chapter, a comprehensive comparison study, including a detailed review and error analysis of common particle mixing techniques and analytical models, is presented to ensure precise design for a high-

performance artificial RF substrate. The correct analytical model is used in the design and manufacturing of the proposed high-performance RF substrate. Moreover, for the first time, the impact of the excitation direction and the polarization of the incident field on the ε_{eff} of the mixing particle are thoroughly examined and discussed. To verify the analysis, the proposed artificial dielectric-based RF substrate is manufactured and tested in both microwave frequency (S-band) at 3.5 GHz and mm-Wave frequency (W-band) at 77 GHz. The measurements indicate a 45% decrease in ε_{eff} and a 38% reduction in loss $\tan\delta$ within the S-band. In the mm-Wave frequency band, there is a 32% decrease in ε_{eff} and a substantial 72% reduction in $\tan\delta$. An excellent agreement is obtained between the simulated, calculated, and measured results.

Chapter 5 applies the particle mixing theory to the design of a 3D-printed artificial dielectric GRIN lens antenna, offering an interesting and viable solution to address the wide beamwidth issue observed in commercial ridged horn antennas when used for radar characterization with UAVs. The integration of the proposed GRIN-ADLA yielded symmetrical radiation patterns in the principal E- and H-planes, significantly reducing the beamwidth from $120^\circ-40^\circ$ to $35^\circ-20^\circ$ across the desired 3–18 GHz frequency range. This resulted in an average beamwidth reduction of 68%. Additionally, alternative solutions such as hyperbolic and multi-layer stack-up lenses, as well as corrugated ridged horn antennas, are also discussed. The hyperbolic lens exhibited a 50% beamwidth reduction at 3 GHz, while the multi-layer lens and the corrugated horn achieved a reduction of approximately 70% at the same frequency. Importantly, all of these UWB beamwidth reduction solutions meet the beamwidth requirements set for in-situ radar characterization using UAVs.

A “one-size-fits-all” antenna is proposed for radars and communication system

applications overcoming the pattern degradation and wide beamwidth limitations of the current available QRHAs. This is discussed in detail in Chapter 6. The proposed design exhibits outstanding bandwidth performance, by covering a wide frequency range from 1 to 32 GHz. Additionally, it maintains an impressive return loss exceeding 10 dB throughout the entire frequency spectrum. Furthermore, the proposed QRHA maintains beamwidth stability and shows no pattern degradation even at higher frequencies while also achieving good VSWR performance, making it an ideal candidate for a diverse set of applications, including EMC testing and radar calibration.

7.2 Research Outcomes

The following are the major outcomes originated from this research work:

1. This research work introduces, for the first time, a comprehensive theory, a correct analytical model, and a set of guidelines for modeling and manufacturing low-loss artificial dielectric materials. These materials have versatile applications in designing high-performance RF substrates for antennas, microwave components, silicon-based on-chip antennas, artificial lenses, and reducing surface-wave losses. This is published in [19].
2. This study validates the claim related to the incompatibility of commercially available QRHA probes for UAV-based in-situ characterization of radars due to their wide beamwidth performance, especially in the lower part of the frequency band. This study contributed to [21–23].
3. To solve the wide beamwidth problem of the commercially available QRHAs for UAV-based in-situ characterization of radars, this work, for the first time,

proposed several solutions including the design of a 3D-printed graded index (GRIN) dielectric lens [24], multi-layer superstrate lens [25], hyperbolic lens [26], and corrugated QRHA [21].

4. Finally, this dissertation discusses the design and modeling of a “one-size-fits-all” low-cost and simple UWB antenna covering a 1-32 GHz band, to satisfy the requirements of both UWB communication and radar systems operating in the L-, S-, C-, X-, Ku-, and K-bands. The proposed QRHA achieves constant beamwidth along with maintaining good matching performance in the entire frequency band and has no pattern degradation. This work contributes to [27, 28].

7.3 Future Research

The following research directions are proposed based on this work:

- There is a non-existence of rules, guidelines, and procedures for designing high-performance artificial dielectric lenses or materials using Maxwell-Garnett theory based on particles other than air cylinders which are used in this work. Compared to the current technique presented, this can lead to the design and manufacturing of artificial lenses or materials with more freedom and tailored ε_r or $\tan(\delta)$ values based on the requirements.
- Using an artificial dielectric material with a high ε_r value can reduce the thickness of the lens thus lowering the size and weight constraints. The GRIN lens discussed in Chapter 5, for the beamwidth reduction problem, has a thickness that can be reduced if an artificial material is designed having a high ε_r .

- Currently, a stack-up of two planar GRIN lenses is used in literature for 2D beam scanning. It will be interesting to look into the comparison and thus fabrication of a spherical Luneburg lens with a graded index profile. A spherical structure compared to the planar lens discussed in this dissertation, is ideal for beam scanning in both planes.
- Currently, QRHAs require substantial computational resources for their full characterization. An analytical model will be useful to quickly evaluate their performance in the first phase of the design before delving into any simulation.

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Appendix A

List of Acronyms and Abbreviations

UWB	Ultrawideband
WISM	Wideband Instrument for Snow Measurements
EMC	Electromagnetic Compatibility
SAR	Synthetic Aperture Radar
TADM	Tunable Artificial Dielectric Material
mm-Wave	Millimeter Wave
RF	Radio Frequency
GRIN	Graded-index
ADLA	Artificial Dielectric Lens Antenna
QRHA	Quad-ridged Horn Antenna
DRHA	Dual-ridged Horn Antenna
GPR	Ground-penetrating Radar
SATCOM	Satellite Communications

SDR	Software-defined Radio
GPS	Global Positioning System
UAV	Unmanned Aerial Vehicle
SS	Stop Sign
HPBW	Half-power Beamwidth
VSWR	Voltage Standing Wave Ratio
TCDA	Tightly-Coupled Dipole Array
PUMA	Planar UWB Modular Antenna
PCB	Printed Circuit Board
PTCDA	Planar Tightly-coupled Dipole Array
TSA	Tapered-slot Array
BAVA	Balanced Antipodal Vivaldi Antenna
AVA	Antipodal Vivaldi Antenna
FIA	Frequency Independent Antenna
CNC	Computer Numerical Control
DMLS	Direct Metal Laser Sintering
PAR	Phased Array Radar
STSR	Simultaneously Transmit and Simultaneously Receive
ATSR	Alternatively Transmit and Simultaneously Receive

ATAR	Alternatively Transmit and Alternatively Receive
AUT	Antenna Under Test
SNR	Signal-to-noise Ratio
ARRC	Advanced Radar Research Center
V-polarized	Vertical Polarization
H-polarized	Horizontal Polarization
CTE	Coefficient of Thermal Expansion
PTFE	Polytetrafluoroethylene
MoM	Method of Moments
HFSS	High-frequency Structure Simulator
SiGe	Silicon-germanium
AoC	Antenna-on-chip
AiP	Antenna-in-package
LCP	Liquid Crystal Polymer
GND	Ground
HIS	High Impedance Surface
MIMO	Multiple-input-multiple-output
M-MIMO	Massive Multiple-input-multiple-output
UM-MIMO	Ultra Massive Multiple-input-multiple-output

TLS	Tunable Lens Surface
PC-MPA	Proximity-coupled Microstrip Patch Antenna
IE	Integral Equation
MDLA	Multi-layer Dielectric Lens Antenna
HPC	High-performance Computing

Appendix B

RF Connectors

B.1 Power Handling

Figure B.1 shows the power handling capability for commonly used RF connectors. A 3.5 mm connector is used for QRHA simulations discussed in Chapter. 6. It can be seen that at the maximum frequency of 34 GHz, the 3.5 mm connector can handle a maximum input power of 45 Watts.

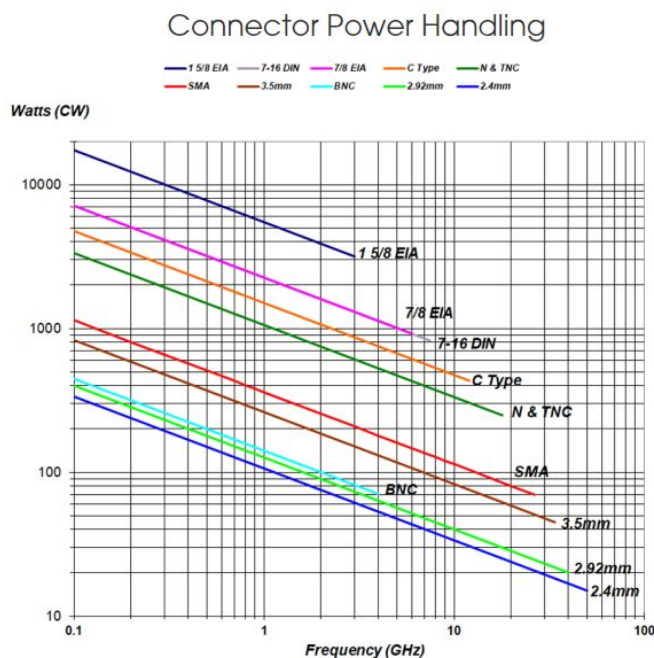


Figure B.1: Power handling chart for commonly used RF connectors [176].

B.2 3.5 mm Female Precision Connector

Figure B.2 shows detailed dimensions of the 3.5 mm female precision connector. This connector has overall dimensions of 9.53 mm $\times$ 9.53 mm $\times$ 13.3 mm and has an air-dielectric interface. The central female connector jacket accepts a pin of 0.92 mm in diameter.

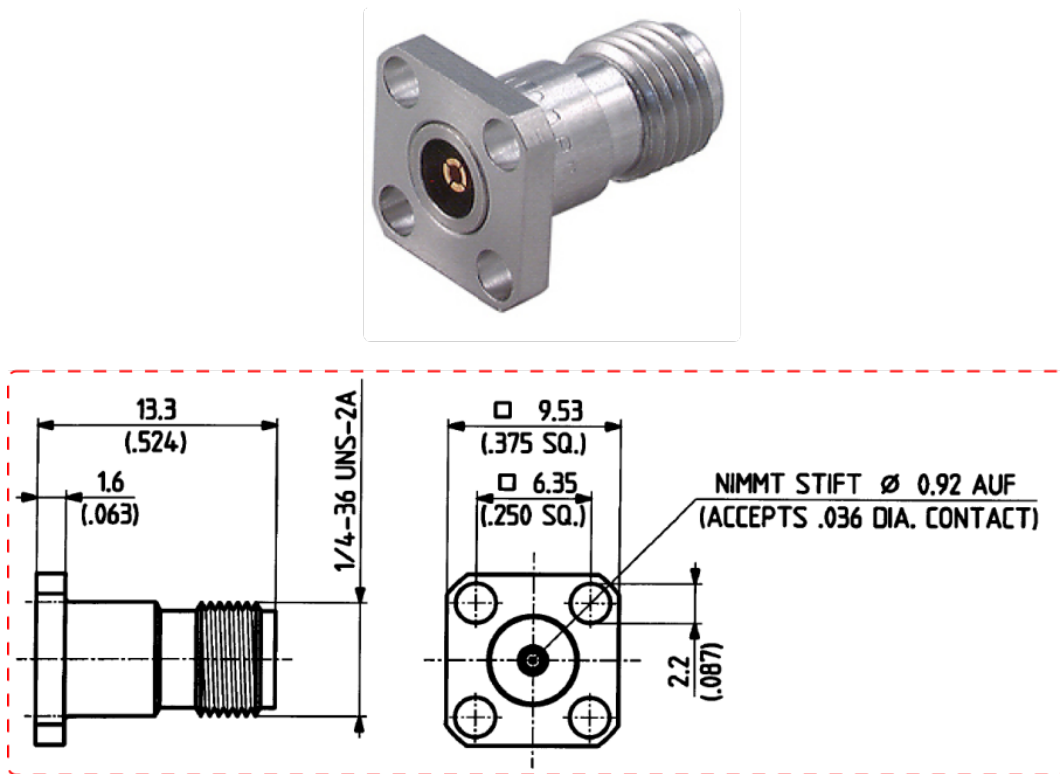


Figure B.2: 3.5 mm female precision connector detailed dimensions.

Appendix C

QRHA CNC Fabrication Preparation

Figure C.1 shows the selection of various parameters such as surface roughness (R_a), fabrication tolerance, and type of aluminum to be used for fabricating the proposed QRHA. For illustration, this figure also shows the comparison of finished and unfinished surfaces of a fabricated part.

Material	Aluminum 6061-T6
Treatment	MIL-DTL-5541, type II, Class 3 Equivalent to Alodine 2000, MIL-C-5541, Class 3
Connector	PC 3.5 Female: (Huber+Suhner: type 23 PC35-50-0-51/199UE)
Roughness (R_a)	32-125 μ in $R_a = 125 \mu$ in
Precision (mm)	± 0.005 in or 0.127 mm

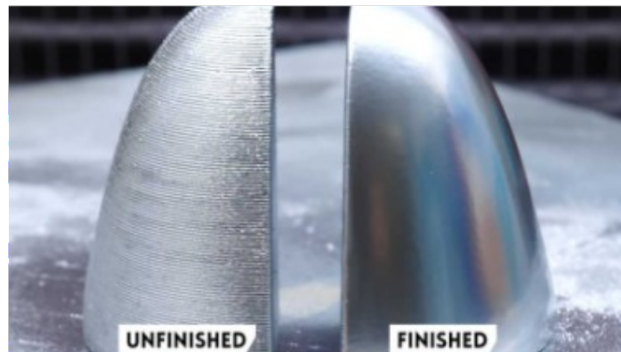


Figure C.1: Summary of various parameters for CNC fabrication as well as illustration of finished and unfinished surfaces (source: www.xometry.com).

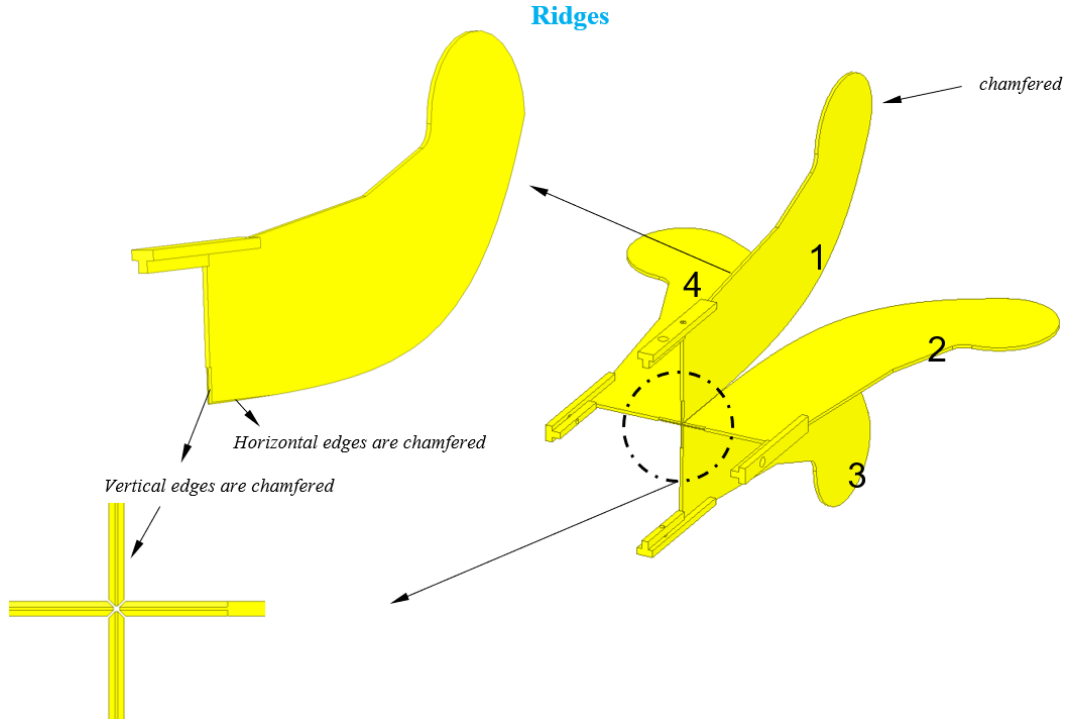


Figure C.2: View of ridges and its chamfered edges.

Figure C.2 shows the individual as well as a complete assembly of four ridges as well as their chamfered regions. Figure C.3 shows the zoomed-in view of the feeding section of the QRHA. It can be seen that the 3.5 mm connector is attached to ridge-1 and has a feeding pin having a diameter of 0.5 mm, clearance of 0.8 mm, and is shorted to the opposite ridge marked as 3. Likewise, for the second polarization, this is repeated for ridge-2 and ridge-4. Figure C.4 shows the integration of ridges, supporting frame, and the cavity. Figure C.5 shows various views of the cavity section including the M3 screws for mounting the supporting frame as well as the four ridges. Figure C.6 and C.7 show the supporting frame with details for various posts and press-fit dowel pins.

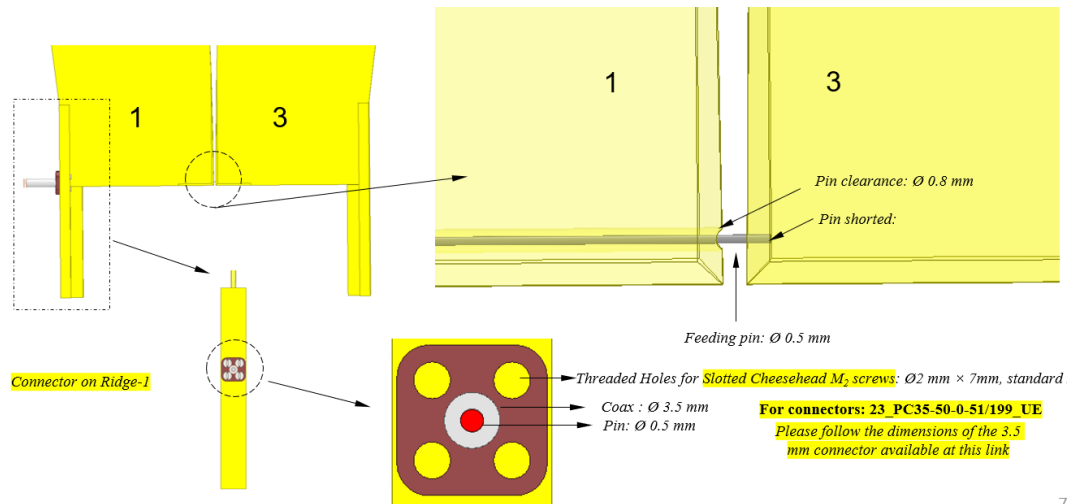


Figure C.3: Close view of the feeding section and the feeding pin.

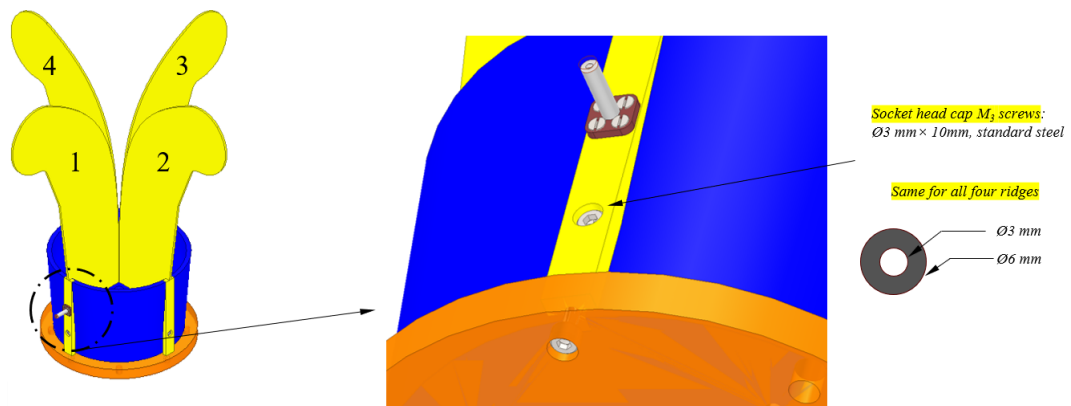


Figure C.4: View of ridges and the required socket head cap M3 screws.

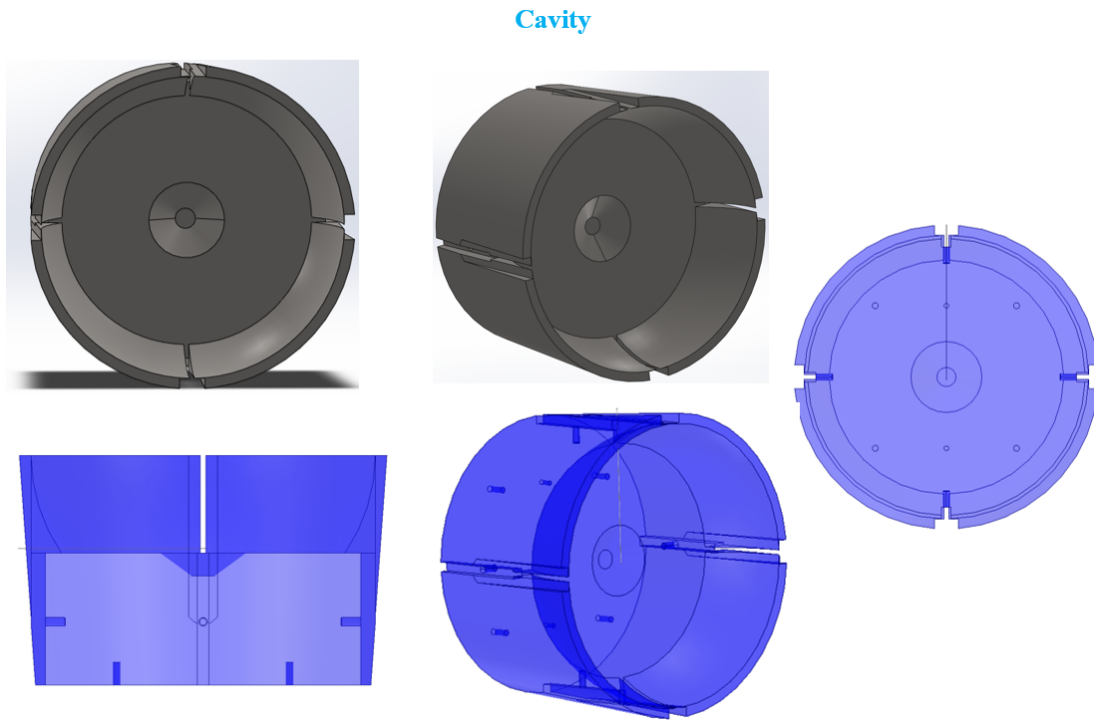


Figure C.5: Different views of the cavity section containing holes for M3 screws.

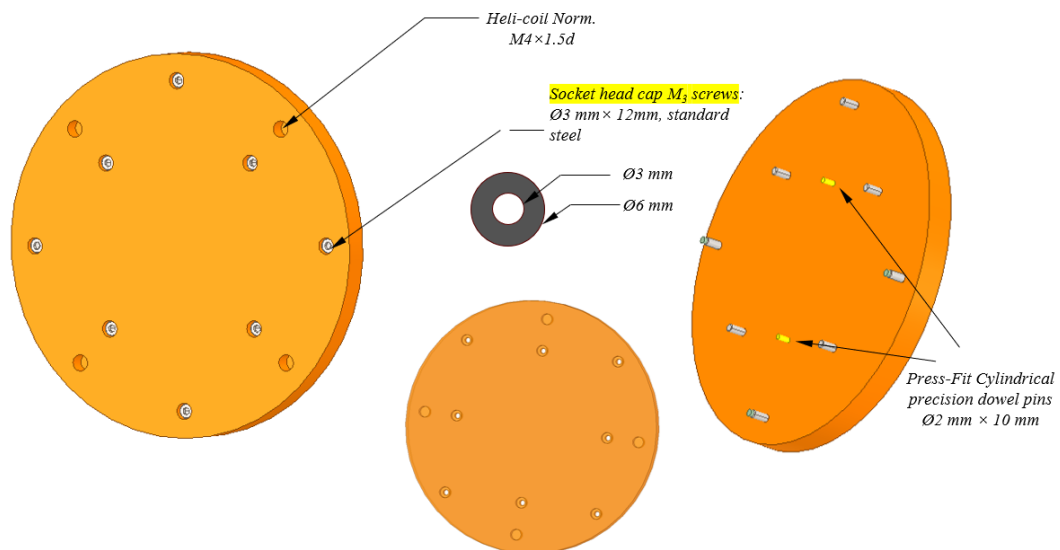


Figure C.6: Supporting frame with screws and cylindrical press-fit dowel pins.

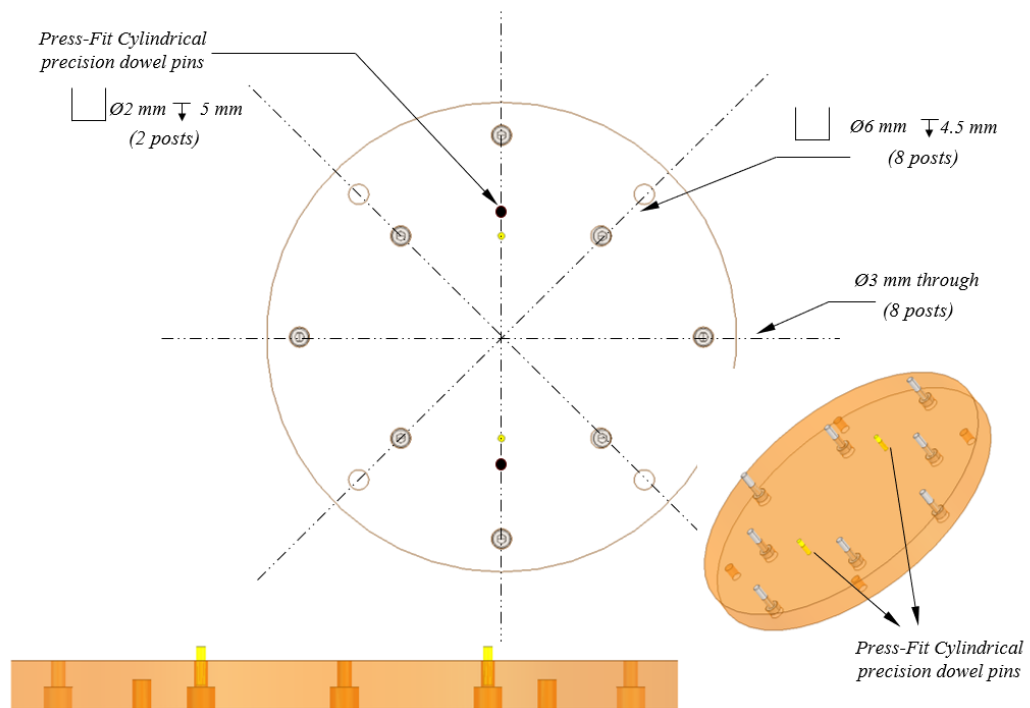


Figure C.7: Another view of supporting frame with screws and cylindrical press-fit dowel pins.

Appendix D

List of Publications

D.1 Patents

- [1] **Syed. S. Jehangir** and J. L. Salazar-Cerreno, “A One-Size-Fits-All Antenna for Radars and Communication Systems,” *US Patent*, in-progress, Oct., 2023

D.2 Journal Papers

- [1] **Syed. S. Jehangir** and J. L. Salazar-Cereno, “A Dual-polarized Ultra-wideband Quad-Ridged-Horn Conical Horn Antenna for Radars and Communication Systems,” *IEEE Transactions on Antennas and Propagation*, in-progress., Oct., 2023.
- [2] **Syed. S. Jehangir**, Sergio Masamura, and J. L. Salazar-Cereno, “An Ultrawideband Artificial Dielectric Lens-Integrated Antenna System with Constant Beamwidth for Radars and Communication Systems,” *IEEE Transactions on Antennas and Propagation*, submitted, July, 2023.
- [3] **Syed. S. Jehangir**, Z. Qamar, N. Aboserwal, and J. L. Salazar-Cereno, “Application of the Mixing Theory in the Design of a High Performance

Dielectric Substrate for Microwave and Mm-Wave Systems,” *IEEE Access*, vol. 8, pp. 180855-180868, 2020, Sep., 2020.

D.3 Conference Papers

- [1] **Syed. S. Jehangir** and J. L. Salazar-Cereno, “Breaking the Limits: A Novel Dual-Polarized Ultrawideband Probe Antenna for High Performance Radar and Communication Systems,” *in the IEEE 45th Annual Meeting and Symposium of the Antenna Measurement Techniques Association (AMTA)*, Seattle, Oct., 2023
- [2] **Syed. S. Jehangir** and J. L. Salazar-Cereno, “An UWB Hyperbolic Lens-Integrated Dual-Polarized Quad-Ridged Horn Antenna for UAV Based In-situ Characterization,” *IEEE International Symposium on Antennas and Propagation (APSURSI)*, Oregon, USA, July, 2023.
- [3] **Syed. S. Jehangir** and J. L. Salazar-Cereno, “The Need for Narrow Beamwidth in Ridged Horn Antennas for UAV-Based In-situ Measurements of Radars and Communication Systems,” *IEEE International Symposium on Antennas and Propagation (APSURSI)*, Colorado, USA, July, 2022.
- [4] **Syed. S. Jehangir** and J. L. Salazar-Cereno, “Achieving Near-Constant Beamwidth and Symmetry in Patterns of the Pyramidal Ridged Horn Antenna for UAV-Based In-situ Characterization and Measurement of Phased Array Radars,” *IEEE Texas Symposium on Wireless & Microwave Circuits and Systems*, Waco, Tx, April, 2022.
- [5] J. L. Salazar-Cereno, **Syed. S. Jehangir**, Antony Segales, Z. Qamar, and N. Aboserwal, “A UAV-based Polarimetric Antenna Measurements for Radar

and Communication Systems from 3-32 GHz,” *IEEE Conference on Antenna Measurements & Applications (CAMA)*, France, pp. 55-60, Oct, 2021.

- [6] J. L. Salazar-Cereno, **Syed. S. Jehangir**, Antony Segales, Z. Qamar, and N. Aboserwal, “An Ultrawideband UAV-based Metrology Platform for In-situ EM Testing of Antennas, Radars, and Communication Systems,” *IEEE Radar Conference*, New York, USA, pp. 1-5, March, 2022

Appendix E

Ph.D. Competition Awards

[1] **Ph.D. Dissertation, Gallogly College of Engineering Dissertation Excellence Award, Oct., 2023.**

[2] **2nd place and a cash prize in the Poster Competition**

“Designing Ultrawideband High Precision Dual-polarized Antenna Probes for Unmanned Aerial Vehicle (UAV) Based Real-time Calibration of Digital Phased Array Radars for Defense and Meteorology Applications”, Oklahoma Aerospace & Defense Innovation Institute (OADII) Student Poster Competition, Oct., 2022.

[3] **1st place in IEEE Industrial Engineering Paper Award**

J. L. Salazar-Cereno, **Syed. S. Jehangir**, Antony Segales, Z. Qamar, and N. Aboserwal, “A UAV-based Polarimetric Antenna Measurements for Radar and Communication Systems from 3-32 GHz,” *IEEE Conference on Antenna Measurements & Applications (CAMA)*, France, pp. 55-60, Oct, 2021.