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Dielectric based antennas for sub-THz applications

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Mémoire présenté en vue de l'obtention du diplôme de *Maîtrise ès sciences appliquées*

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Dielectric based antennas for sub-THz applications

présenté par **Ehsan RAHMATI**

en vue de l'obtention du diplôme de *Maîtrise ès sciences appliquées*

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DEDICATION

To my parents and my love, Paris.

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I would like to express my profound gratitude to my supervisor, Prof. Mohammad S. Sharawi, for his priceless advice, inspiration, patience, and support. I am extremely appreciative of the chance he gave me to work with his research team as well as the inspiration and guidance he provided for me to finish this project.

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Last but not least, I would like to thank my parents and Paris, my love, from the bottom of my heart for their motivation and support.

RÉSUMÉ

Ces dernières années, l'augmentation du trafic de données sans fil et le nombre croissant de nouvelles applications a souligné l'importance d'avoir de larges bandes de fréquences dans les bandes Sub-THz et THz pour l'avenir des communications. Ce sera primordial en tant que technologie viable dans la 6G et au-delà pour atteindre des niveaux élevés de débits de données. Cependant, la perte de trajet et l'atténuation atmosphérique sont élevées, ce qui limite considérablement les communications sans fil et, par conséquent, des antennes à gain élevé et à large bande sont nécessaires. Donc, avec l'avancement de la technologie de fabrication, la conception d'antennes fonctionnant à sub-THz fréquence peut être une solution potentielle pour la difficulté considérable de trouver des spectres pour des applications telles que les applications de réseau local point à point fixe et sans fil.

Dans cette thèse, tout d'abord, un nouveau guide d'onde diélectrique intégré comme mécanisme d'alimentation pour le diélectrique des antennes à base sont présentées et fabriquées dans la gamme de fréquences de 220-280 GHz à l'aide du laser technologie de micro-usinage. La structure conçue fonctionne sur la gamme de fréquences de 220-280 GHz avec la perte d'insertion appropriée entre -2 dB et -4 dB. Puis, roman à deux éléments antennes à base diélectrique dans la gamme de fréquences sub-THz (230-270 GHz) pour les normes sans fil qui sont performants et facilement intégrables avec des appareils électroniques sont présentés. Analyse de ces structures l'utilisation d'outils de simulation numérique basés sur des techniques numériques EM indique le rayonnement élevé efficacité (moyenne de 93,6 %) et gain (moyenne de 13,77 dBi) avec les diagrammes de rayonnement latéraux.

Enfin, la nouvelle antenne Sub-THz est fabriquée pour comparer et valider les caractéristiques du structure proposée avec les résultats de la simulation. De plus, il a été testé avec succès. L'avantage de notre projet de recherche peut être la présentation d'une nouvelle antenne à haute diélectrique pour les applications de communication sans fil dans la bande de fréquence Sub-THz.

ABSTRACT

In recent years, growing wireless data traffic and an increasing number of new applications have highlighted the importance of having wide frequency bands in the Sub-THz and THz bands for future communications. This will be paramount as a viable technology in 6G and beyond to achieve high data rates. However, the path loss and atmosphere attenuation are high, which significantly limits wireless communications, and as a result, high gain and wideband antennas are needed. Therefore, with the advancement of fabrication technology, the design of antennas operating at a Sub-THz frequency can be a potential solution without the considerable difficulty of finding unallocated spectrums for applications, such as fixed point to point and wireless local area network applications.

In this thesis, first, a new integrated dielectric waveguide as a feeding mechanism for the dielectric based antennas is presented and fabricated in the frequency range of 220-280 GHz using the laser micromachining technology. The designed structure operates in the frequency range of 220-280 GHz with the appropriate insertion loss between -2 dB and -4 dB. Then, novel dual-element, dielectric based antennas at the Sub-THz frequency range (230-270 GHz) for wireless standards that are efficient and easily integrable with electronic devices are presented. Analysis of these structures using numerical simulation tools based on EM numerical techniques indicates the high radiation efficiency (average of 93.6%) and gain (average of 13.77 dBi) with the broadside radiation patterns.

Finally, the novel Sub-THz antenna is built to compare and validate the characteristics of the proposed structure with the simulation results. Moreover, it has been successfully tested. The benefit of our research project is the presentation of a newly designed, high gain, dielectric-based antenna for wireless communication applications in the Sub-THz frequency band.

TABLE OF CONTENTS

DEDICATION	III
ACKNOWLEDGEMENTS	IV
RÉSUMÉ.....	V
ABSTRACT	VI
TABLE OF CONTENTS	VII
LIST OF TABLES	X
LIST OF FIGURES.....	XI
LISTE OF SYMBOLS AND ABBREVIATIONS	XVI
LIST OF APPENDICES	XVIII
CHAPTER 1 INTRODUCTION	1
1.1 Wireless Communication Systems	1
1.2 Motivation	4
1.3 Objectives	5
1.4 Methodology	6
1.5 Thesis Organization.....	7
CHAPTER 2 BACKGROUND AND LITERATURE REVIEW	8
2.1 Theory and Fundamental of Dielectric Resonator Antennas	8
2.1.1 Cylindrical Dielectric Resonator Antenna	9
2.1.1.1 Analysis of the First Order Model of the Cylindrical Dielectric Resonator Antenna (CDRA)	9
2.1.1.2 Design Procedure of a Cylindrical DRA	16
2.1.2 Rectangular Dielectric Resonator Antenna	18
2.2 Feeding Techniques of DRAs	19

2.2.1	Array feed networks	22
2.3	Bandwidth Enhancement Techniques	22
2.4	Conclusion.....	24
CHAPTER 3 DIELECTRIC DRA DESIGN FED BY SUBSTRATE INTEGRATED WAVEGUIDE (SIW) AT MILLIMETER WAVE BAND		25
3.1	Design of the Substrate Integrated Waveguide as the Feeding Structure of DRA	25
3.2	Design of SIW Fed-Cylindrical Dielectric Resonator Antenna	27
3.3	Design of Perforated Dielectric Resonator Antenna	29
3.4	Design of 1×4 Substrate Integrated Waveguide -Power Divider	32
3.5	Design of SIW Parallel Fed-Cylindrical Dielectric Resonator Antennas Array	33
3.6	Conclusion.....	36
CHAPTER 4 DESIGN OF NOVEL DIELECTRIC BASED ANTENNAS INTEGRATED WITH LOW LOSS FEEDING SUBSTRATE INTEGRATED GUIDES.....		37
4.1	Introduction	37
4.2	Designed Antennas Fed by Planar Dielectric Waveguides	37
4.3	Design of a Feeding Substrate-Integrated Guide in the Sub-THz Band	41
4.3.1	Design of a Substrate-Integrated Silicon Waveguide Supported by a Pyrex Layer at an Operating Frequency Range of 140-180 GHz.....	41
4.3.2	Design of a Substrate-Integrated Alumina Waveguide Supported at an Operating Frequency Range of 220-280 GHz.....	44
4.1.1.1	Simulation and Discussion	44
4.1.1.2	Fabrication Design	46
4.1.1.3	Measurement Results and Discussion	47
4.4	Design of Novel DBAs Integrated with Proposed Low Loss Feeding Substrate Integrated Guide in the Sub-THz Band.....	49

4.4.1	Design of a Novel Sub-THz Dielectric Resonator Antenna Fed by Substrate Integrated Image Guide (SIIG) Technology at 165 GHz	50
4.4.2	Design of a Novel Dielectric Based Antenna Fed by the Proposed Feeding Technique in the Frequency Range of 230-270 GHz.....	53
4.1.1.4	Simulation and Discussion	54
4.1.1.5	Fabrication.....	57
4.1.1.6	Fabrication Errors	59
4.1.1.7	Measurement Results and Discussion	60
4.5	Conclusion.....	65
CHAPTER 5 CONCLUSION AND FUTURE WORKS.....		67
5.1	Conclusion.....	67
5.2	Future Works.....	68
REFERENCES.....		69
APPENDICES		78

LIST OF TABLES

Table 1.1	ITU assigned bands in the 100–275 GHz frequency range, Reproduced with permission from Springer Nature [18]	3
Table 2.1	Advantages and disadvantages of DRAs	9
Table 2.2	The dimensions of optimized microstrip-fed aperture DRA parameters	21
Table 3.1	The dimension size (in mm) of optimized SIW-DRA parameters.....	28
Table 4.1	The parameter dimensions (in mm) of the designed DRA	51
Table 4.2	A comparison summary between previous works and the proposed DBA.....	65

LIST OF FIGURES

Figure 1.1	The evolution of mobile wireless generation [2]	1
Figure 1.2	Spectrum bands of 4G, 5G, and 6G [17]	2
Figure 1.3	A variety of use cases will be enabled by 6G systems [6].....	3
Figure 1.4	Atmospheric absorption across mm-wave frequencies in dB/km © [2011] IEEE [19]	4
Figure 2.1	Different shapes of dielectric resonator, with permission from publisher [32]	8
Figure 2.2	Field distributions of different modes of CDRA, (a) H field distribution of $HEM_{11\delta}$ mode (b) E field distribution of $TE_{01\delta}$ mode, and (c) H field distribution of $TM_{01\delta}$ mode © [1984] IEEE [39]	10
Figure 2.3	First order model of the dielectric resonator cavity	11
Figure 2.4	Geometry of the cylindrical DRA.....	16
Figure 2.5	Calculated K_{0a} vs (a/h) for cylindrical DRA designs at 64 GHz for $\epsilon_r = 10$	17
Figure 2.6	Geometry of the rectangular DRA.....	18
Figure 2.7	(a) Geometry of the microstrip-fed aperture © [2011] IEEE [43], and (b) Simulated H field distribution using HFSS tool v2021	19
Figure 2.8	(a) $HEM_{11\delta}$ mode of a CDRA by the probe feeding technique © [2015] IEEE [44], (b) Rectangular DRA array prototype © [2018] IEEE [46].....	20
Figure 2.9	(a) Geometry of the CDRA, and (b) obtained S_{11} parameter.....	20
Figure 2.10	(a) Fabricated prototype series-fed SIW DRA © [2011] IEEE [50], (b) Geometry of the DRA array © [2018] IEEE [53]	21
Figure 2.11	Array feeding networks. (a) Parallel feed and (b) Series feed	22
Figure 2.12	Photographs of prototypes (a) SIDRA © [2021] IEEE [61], (b) Hybrid-glass DRAs [63]	23
Figure 2.13	Configuration of the designed DRA using 3D printing technology © [2019] IEEE [69]	23
Figure 3.1	Geometry of the SIW © [2005] IEEE [73]	25

Figure 3.2 (a) Geometry of the designed SIW, (b) Magnitude of the electric field distribution at frequency of 64 GHz	27
Figure 3.3 S parameters of the designed SIW	27
Figure 3.4 (a) SIW- based CDRA driven model, (b) The impact of slot position on the reflection coefficient	28
Figure 3.5 (a) The impact of DRA radius on the reflection coefficient, (b) radiation pattern of the CDRA at frequency of 66 GHz	29
Figure 3.6 The geometry of the proposed structure (a) Isometric view (b) Top view	30
Figure 3.7 The simulated S_{11} parameter of the proposed perforated DRA	31
Figure 3.8 (a) The gain of the proposed antenna at the boresight direction ($\theta = 0^\circ$), and (b) The Radiation pattern (gain) of the proposed perforated DRA at the resonance frequency of 60 GHz	32
Figure 3.9 (a) Structure of a 1×4 (6 dB) SIW-power divider, (b) E-field distribution at a frequency of 64 GHz	33
Figure 3.10 S parameters (dB) of the designed SIW based power divider	33
Figure 3.11 Structure of the SIW Parallel-Fed CDRA (a) Top view (b) Isometric view	34
Figure 3.12 (a) Simulated reflection coefficient of parallel array configurations for the longitudinal slot position, (b) radiation pattern of the CDRA at a resonance frequency of 63.75 GHz	34
Figure 3.13 (a) Structure of SIW parallel fed-CDRAs with transition, (b) The simulated reflection coefficient of parallel fed CDRA array with MSL transition	35
Figure 3.14 Radiation pattern (gain) of parallel fed array with MSL transition at resonance frequencies of (a) 61.8 GHz and (b) 64 GHz	35
Figure 4.1 (a) Tapered SOG antenna, (b) Fabricated sample, and (c) Measured and simulated S_{11} parameters © [2014] IEEE [95]	38
Figure 4.2 (a) Fabricated SIIG rod antenna at 94 GHz (b) E-plane radiation pattern in $\varphi = 90^\circ$ © [2006] IEEE [96]	38

Figure 4.3 (a) Geometry of the 7-element DRA fed by DIG (b) Simulated and measured S_{11} parameter © [2010] IEEE [97], (c) Geometry of the 7-element DRA coupled to DIG through aperture slots with back reflector, (d) Simulated and measured S_{11} parameter © [2009] IEEE [98].....	39
Figure 4.4 (a) Fabricated structure from a single silicon wafer, (b) and (c) Simulated and measured radiation patterns at 325 GHz in the xz plane and yz plane, respectively [99]	40
Figure 4.5 (a) Geometry of the DRA based on topological photonic crystals, (b) Top view, (c) Reflection coefficient, and (d) Gain and radiation efficiency, with the permission of AIP Publishing [101].....	40
Figure 4.6 The 3D view of the proposed substrate integrated waveguide (a) Top view, (b) Side view, and (c) Perspective view	42
Figure 4.7 (a) S_{11} parameter and (b) S_{21} parameter (in dB) of the proposed substrate integrated waveguide for different values of HR-silicon conductivity	42
Figure 4.8 Magnitude of the electric field distribution at 160 GHz.....	43
Figure 4.9 (a) S_{11} parameter and (b) S_{21} parameter (in dB) of the proposed substrate integrated waveguide for different air gap thicknesses	43
Figure 4.10 (a) Top view of the proposed substrate integrated guide (b) Side view	44
Figure 4.11 (a) Simulated reflection and transmission coefficients of the proposed waveguide as a function of frequency for different values of (a) guiding channel width (w_c) when the waveguide thickness is $t_s=0.25$ mm and (b) waveguide thickness (t_s) when the guiding channel width is $w_c=0.3$ mm	45
Figure 4.12 (a) Simulated reflection and transmission coefficients of the proposed waveguide as a function of frequency for different values of (a) air gap thickness (t_{airgap}) and (b) air gap width (w_{airgap}) when the guiding channel width and waveguide thickness are $t_s=0.25$ mm and $w_c=0.3$ mm	46
Figure 4.13 The designed short waveguide for the fabrication (a) Top view (b) Diametric view (c) Photograph of the fabricated Alumina waveguide, and (d) Photograph of the presented back to back DBA architecture	47

Figure 4.14 (a) THz measurement set-up (b) the proposed waveguide under test (DUT).....	48
Figure 4.15 The simulated and measured S-parameters of the proposed back to back structure with different lengths (a) Short waveguide (SWG) and (b) Long waveguide (LWG).....	48
Figure 4.16 E field distribution of the (a) guiding channel and (b) HEM_{11} mode of a cylindrical dielectric resonator © [1984] IEEE [39]. H field distribution of the (c) guiding channel and (d) HEM_{11} mode of a cylindrical dielectric resonator © [1984] IEEE [39]	50
Figure 4.17 The geometry of the proposed integrated antenna (a) 3D perspective view and (b) Top view	51
Figure 4.18 (a) Simulated S parameter of the proposed antenna and (b) H-Field distribution inside the DRA in the guiding channel of SIIG at the frequency of 165 GHz	52
Figure 4.19 (a) Gain of the proposed antenna at broadside direction ($\theta=0^\circ$), (b) Radiation pattern (gain) of the proposed antenna at resonance frequency of 165 GHz	53
Figure 4.20 The geometry of the proposed dielectric based antenna (a) Diametric view, (b) 2x1y (top view), and (c) 2x3y (top view).....	54
Figure 4.21 (a) S_{11} of the (2x1y) antenna for different dimensions of r_2 , (b) investigation of the air hole matching on S_{11} of the (2x1y) antenna, (c) S parameters of the designed DBA for two different configurations of (2x1y) and (2x3y), and (d) Active S parameters of the designed DBA for two different configurations of (2x1y) and (2x3y)	55
Figure 4.22 (a) Radiation efficiency and (b) gain (dBi) along the z-axis of two different configurations of (2x1y) and (2x3y) DBA	56
Figure 4.23 Simulated radiation patterns (gain) of the proposed dual-element DBA for the (2x3y) configuration, in the xz-plane (E-plane-blue line) and yz-plane (H-plane-red line) at the frequencies of (a) 237 GHz, (b) 244 GHz, (c) 253 GHz, and (d) 262.5 GHz	57
Figure 4.24 (a) and (b) The designed DBA for the fabrication, (c) and (d) The applied epoxy layers, (e) Photograph of the fabricated alumina layer, and (f) Photograph of the presented back-to- back DBA architecture.....	58
Figure 4.25 (a) and (b) Photographs of the aluminium fixture under testing with and without the DRBA, respectively. (c) and (d) Measured S-parameters for the aluminium fixture with	

and without the DRBA, respectively. (e) and (f) Simulated and measured S-parameters for the aluminium fixture without DBA, respectively	59
Figure 4.26 (a) THz measurement set-up (b)Top view of the proposed DBA under testing (DUT)	60
Figure 4.27 Simulated and measured (a) the reflection coefficient and (b) the mutual coupling for the proposed antenna	61
Figure 4.28 (a), (b), and (c) Photographs of the anechoic chamber and measurement set-up	62
Figure 4.29 The simulation and measured radiation patterns of the designed DBA for three different frequencies of (a) and (b) 247 GHz, (c) and (d) 257 GHz, (e) and (f) 267 GHz in the xz-plane and yz-plane.....	63
Figure 4.30 Simulated and measured radiation patterns (combined gain) of the proposed dual-element DBA for the (2x3y) configuration at three various frequencies of (a) and (b) 247 GHz, (c) and (d) 257 GHz, (e) and (f) 267 GHz in the xz-plane and yz-plane...	64

LISTE OF SYMBOLS AND ABBREVIATIONS

4G	Fourth Generation
5G	Fifth Generation
AI	Artificial Intelligence
BW	Bandwidth
CST	Computer Simulation Technology
CMOS	Complementary Metal-Oxide Semiconductor
DRA	Dielectric Resonator Antenna
DBA	Dielectric Based Antenna
DIG	Dielectric Image Guide
FCC	Federal Communication Commission
HFSS	High Frequency Structure Simulator
HR-Si	High-Resistivity Silicon
ITU	International Telecommunication Union
IoE	Internet of Everything
MSL	Microstrip Line
PCB	Printed Circuit Board
PEC	Perfect Electric Conductor
PMC	Perfect Magnetic Conductor
SHA	Standard Horn Antenna
SIIG	Substrate Insular Image Guide
SIDRA	Substrate-Integrated Dielectric Resonator Antenna
SIIG	Substrate Integrated Image Guide

SIW	Substrate Integrated Waveguide
VNA	Vector Network Analyzer

LIST OF APPENDICES

APPENDIX A LIST OF PUBLICATIONS	78
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CHAPTER 1 INTRODUCTION

1.1 Wireless Communication Systems

The last two decades have seen an increase in interest in wireless communication. The first generation wireless system, based on analog transmission, was introduced as the starting point. However, this technology had some limitations, for instance, security problems [1]. These restrictions were removed by the second generation to some extent. Video calls became possible in 2000 using third generation communications networks [2]. Significant investments were made in the communications industry, which led to the development of fourth generation (4G) systems. After that, fifth generation (5G) systems, which can provide faster speeds and have a wider frequency band, were introduced [2-4]. Figure 1.1 depicts the evolution of mobile wireless communications networks from 1G to 6G.

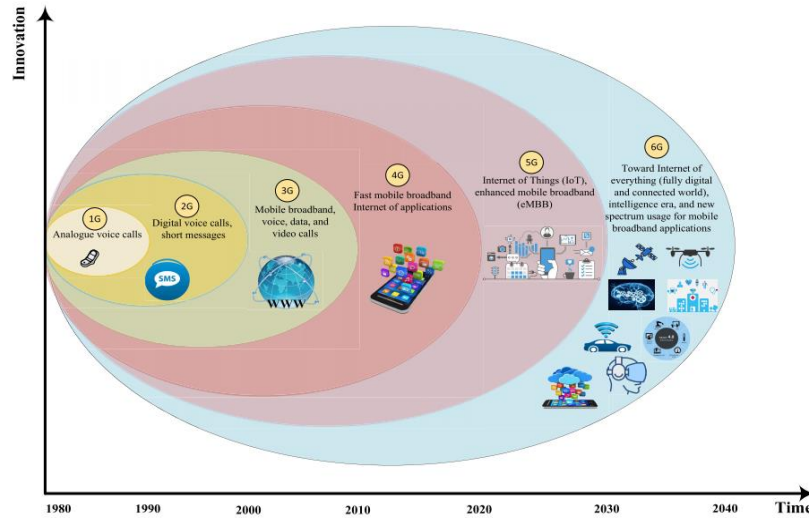


Figure 1.1 The evolution of mobile wireless generation [2]

The need for more efficient communication systems is demonstrated by the high volume of traffic on networks, the growth of emerging applications, like the Internet of Everything (IoE) and artificial intelligence (AI), the required Tbps-level data rates, and a need to improve energy efficiency and low latency [5, 6]. In 2030, it is predicted that the traffic volume of wireless connectivity will exceed 5016 (EB/monthh), and there will be 17.1 billion mobile subscriptions

[5]. Therefore, moving to higher frequency bands is required to get around these constraints, thereby providing an effective response to the increasing demand.

Millimeter-wave (mm-wave 30GHz -300GHz) frequency bands and Sub-THz frequency bands (100-300GHz) have gained much attention. 5G wireless communications operating at the millimeter-wave band can provide a multi-Gbps data rate for mobile devices and new applications in comparison to the sub-6 GHz frequency range [7]. At the millimeter-wave band, some global unlicensed wireless spectrums, 57 to 71 GHz, for instance, that the US Federal Communication Commission (FCC) allocated can be used for 5G applications [8]. The unlicensed spectrum from 57-64 GHz in the millimeter wave band is thus attracting attention for broadband wireless applications within short distances [9-11]. However, because of propagation loss, atmospheric attenuation, and sensitivity to blockage, wireless communications in these frequency ranges differ significantly from those at lower frequencies [12-15]. Therefore, a high gain and wideband antenna in these frequencies is necessary due to the propagation losses and atmospheric absorption [10, 15]. Sub-THz and THz systems are thought to be a promising solution for applications in 6G and beyond, because the necessity for higher frequency bands is imminent and finding unallocated spectrums below 100 GHz is not easy due to their incorporation into the 5G standards [16]. Figure 1.2 shows the spectrum bands of 4G, 5G, and 6G systems. The prominent features of the mmWave and THz systems not only enable autonomous vehicles, robotic controls, high-definition holographic gaming, and high-speed wireless data distribution but also develop applications like sensing, imaging, and wireless communication, to name but a few, in the 6G networks and beyond [14].

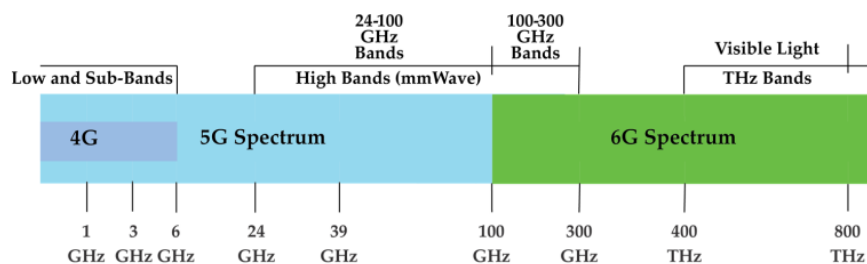


Figure 1.2 Spectrum bands of 4G, 5G, and 6G [17]

A 6G system can provide 1000 times more wireless connectivity than the 5G system [5]. Additionally, in comparison to 5G systems, the larger bandwidth and higher data rate provide more

accurate localization and sensing [4]. Figure 1.3 shows a variety of use cases that will be enabled by 6G systems.

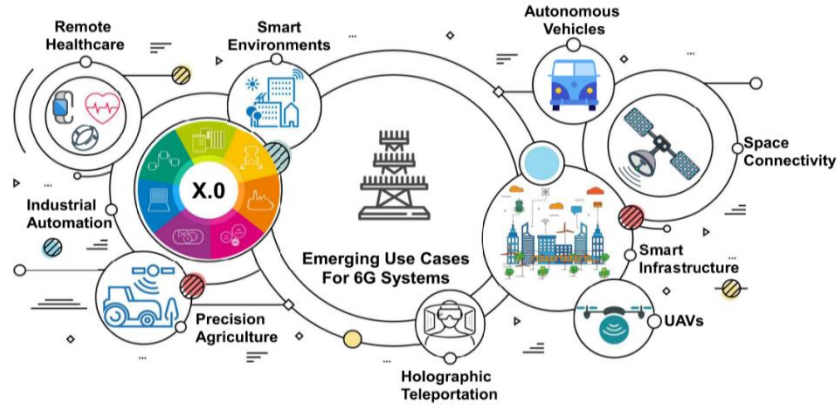


Figure 1.3 A variety of use cases will be enabled by 6G systems [6]

By considering the mentioned features, research teams have been interested in 6G systems for years, and the frequency range of above 95 GHz to 3 THz, allocated by the Federal Communications Commission (FCC), can be used for future 6G research [2]. Table 1.1 shows the allocated bands by the ITU in the frequency range of 100–275 GHz for fixed and mobile use. Consequently, according to the mentioned limitations and advantages, the focus of this research project will be in the Sub-THz frequency band.

Table 1.1 ITU assigned bands in the 100–275 GHz frequency range, Reproduced with permission from Springer Nature [18], [16]

Band (GHz)	Bandwidth (GHz)	Band (GHz)	Bandwidth (GHz)
102–109.5	7.5	167–174.8	7.8
111.8–114.25	2.45	191.8–200	8.2
122.25–123	0.75	209–226	17.0
130–134	4.0	232–235	3.0
141–148.5	7.5	238–241	3.0
151.5–164	12.5	252–275	23.0

1.2 Motivation

In the modern era, the demand for high data rates has been rising along with wireless data traffic in the communications industry. Taking this into consideration, there are key advantages of THz wireless communications, i.e., a large usable frequency band as well as a small antenna, and the advancement of fabrication technology. The design of antennas operating at a Sub-THz frequency could a potential solution without the considerable difficulty of finding unallocated spectrums below 100 GHz for new applications to achieve high data rates [16]. However, wireless communications in millimeter wave and Sub-THz bands are more challenging due to the free space path loss and high atmosphere attenuation [7, 13, 19]. Figure 1.4 shows the atmospheric absorption characteristics of mm-wave propagation [20]. Therefore, to compensate for losses, high gain and highly directional antennas are required [14, 21]. Dielectric resonator antennas proposed by Long et al [22] can be employed due to their advantages such as high gain, low loss, small size, large bandwidth, and high efficiency [23, 24].

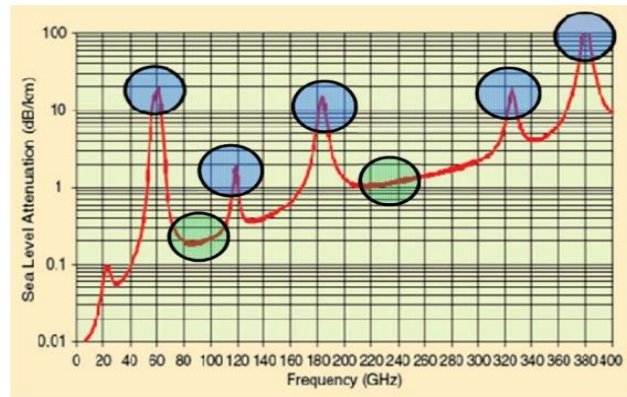


Figure 1.4 Atmospheric absorption across mm-wave frequencies in dB/km © [2011] IEEE [20]

Several feeding techniques have been reported for the exciting of DRA modes, such as aperture coupling [25], coplanar waveguide [26, 27], and probe [28, 29]. However, they cause a significant conductivity loss at high frequencies. Therefore, the need for a feeding mechanism due to the low loss and technology compatibility in the Sub-THz band is inevitable.

The presentation of a low loss feeding substrate with integrated guide and a novel, high gain dielectric-based antenna integrated with low loss feeding dielectric guides that can operate at

various frequency ranges of 140-300 GHz for wireless communications will be an interesting design. Moreover, the design of a high gain Sub-THz dielectric based antenna and antenna arrays based on the technology platforms of silicon-based systems in a package for 6G wireless standards that are efficient and easily integrable with electronic devices is a response to future demands. Additionally, a linear DRAs array fed using the feeding SIW-power divider at the frequency range of 57-64GHz can be an interesting choice for short wireless communication. An analysis of these structures will be performed using numerical simulation tools based on electromagnetic numerical techniques. The novel Sub-THz antenna will be fabricated to compare and validate the characteristics of the proposed structure with the simulation results.

The advantage of our research project is the presentation of newly designed antennas integrated with a low loss feeding substrate integrated guides for wireless communication applications in the Sub-THz frequency band.

1.3 Objectives

The general objective of this thesis is the design of a high gain, small size, wideband, completely embedded antenna in a waveguide system in the Sub-THz band with the capability of integration with other devices in a planar form.

The specific objectives of the research project are:

- 1-Design of a low loss substrate integrated waveguide (SIW) in a millimeter wave frequency band (50-70 GHz), employing it as a feeding mechanism to excite the perforated dielectric based antenna and the cylindrical dielectric resonator antenna array.

- 2-Design, study, and fabrication of a low loss feeding substrate-integrated dielectric waveguide with the capability of integration with other devices in a planar form and a low manufacturing cost, testing the designed prototype, and comparing the obtained simulation results with the measurement results of the fabrication over the frequency range of 220-280 GHz.

3-Design and fabrication of novel completely embedded dual-element, high gain DRAs for wireless communications using a proper directly feeding network at the frequency range of 230-270 GHz and achieving the proper radiation characteristics, namely high gain in the broadside direction, the acceptable impedance matching, and high radiation efficiency.

1.4 Methodology

To achieve the objectives mentioned in the previous section, the methodology of the research project is elaborated below.

After understanding the fundamental concepts of the antennas, their shortcomings, advantages, and disadvantages through comprehensive literature reviews, re-derivation, theoretical investigation of the dielectric resonator modes, and design techniques, the design of the novel DRA begins. Analysis of the structures will be performed using numerical simulation tools based on EM numerical techniques. By doing so, we can compute approximate solutions to Maxwell's equations, thereby predicting structure behaviors. This process will be done for two different structures. First, the design of a DRA array fed by the aperture cut on the broad wall of a SIW power divider at various frequencies between 57-71GHz. Second, the design of a novel high gain dielectric based antenna using a proper feeding mechanism at various frequency ranges in the Sub-THz band (100-300 GHz). Here, the proposed waveguide is examined from a theoretical perspective, and the results are validated using numerical simulation tools. Simulation results are performed using the High Frequency Structure Simulator (HFSS) and CST Microwave Studio tools v2021. Then, the proposed novel Sub-THz antenna will be built to compare and validate the characteristics of the proposed structure with the simulation results.

Additionally, from the perspective of antenna fabrication, there is a challenging point in this research project related to the fabrication limitations and a potential manufacturing error due to the small size of the structure. Therefore, we will have to make sure all the dimensions of the proposed structure are accurate and if the measured results are not correct, we will have to stop and reassess. Integrating all the components with the best-integrated solution for the Sub-THz dielectric-based antenna will be examined via simulations and prototyping. The last step will be the task of characterizing and testing the proposed structures.

1.5 Thesis Organization

This thesis is organized into five chapters. Chapter 1 explains the importance of wireless communications in the Sub-THz band for 6G and beyond. In Chapter 2, a complete literature review of the dielectric resonator antenna, including theory and the fundamentals of dielectric resonator antennas, advantages and disadvantages, and feeding technique, is performed. Dielectric resonator antennas fed by a substrate integrated waveguide (SIW) are presented in Chapter 3. Then, in Chapter 4, the two different substrate-integrated waveguides are presented for frequency ranges of 140-180 GHz and 230-300 GHz. After that, a novel single-element, dielectric resonator antenna integrated with a low loss substrated integrated image guide in the Sub-THz band is proposed. Moreover, the design and fabrication of a novel, dual-element high gain dielectric-based antenna integrated with the proposed low loss feeding mechanism for a frequency range of 240-270 GHz are presented. Finally, Chapter 5 concludes the thesis work, and future works that can be attractive for THz future wireless communications according to the concepts given to this thesis are suggested.

CHAPTER 2 BACKGROUND AND LITERATURE REVIEW

In this chapter, the fundamentals of dielectric resonators, their advantages and disadvantages, the design procedures for cylindrical as well as rectangular DRAs, conventional techniques of increasing the bandwidth, and feeding techniques are presented. Additionally, a review of designed Sub-THz antennas is provided.

2.1 Theory and Fundamental of Dielectric Resonator Antennas

In 1983, the dielectric resonator came into being as an antenna [22]. Due to the elimination of conduction loss, dielectric resonator antennas were a better option than metal-based ones for high frequency applications [30, 31]. During the past decade, DRAs of different shapes have been presented [32-35], as shown in Fig. 2.1, but the conventional shapes of DRAs are hemispherical, cylindrical, and rectangular.

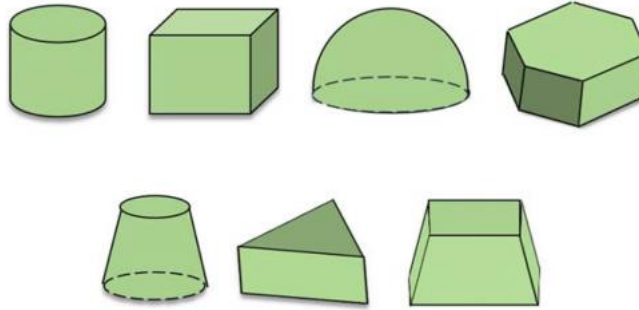


Figure 2.1 Different shapes of dielectric resonator, with permission from publisher [33]

DRAs of different dielectric constants have been made with the advancement of fabrication technology. The size can be scaled up or down by utilizing the dielectric constant value. There is a possibility that DRA excitation will not be effective at high frequencies if a material with a low dielectric constant is chosen to make fabrication easier. Therefore, the trade-off is between fabrication feasibility and the efficient mode excitation. [32]. The dielectric resonator antennas have some advantages and disadvantages, shown below Table 2.1 [23, 36-39], that make them attractive. The variables of λ_0 and ϵ_r are the free space wavelength and relative permittivity of DRAs, respectively.

Table 2.1 Advantages and disadvantages of DRAs

Advantages	Disadvantages
Size scaling ($\propto \lambda_0 / \sqrt{\epsilon_r}$)	Impedance changing by slightly moving from optimum position
No conductor loss	Complex Feeding techniques at Sub-THz band
Large impedance bandwidth	High cost for some building methods
Various excitation techniques	Limitation on fabrication technologies
No surface waves	

In the next sections, the properties and design methods of cylindrical and rectangular DRAs will be discussed.

2.1.1 Cylindrical Dielectric Resonator Antenna

In this section, cylindrical dielectric resonator antenna field configurations (modes) are presented as solutions to satisfy Maxwell's equations along with the boundary conditions. Then, the design procedure for a cylindrical DRA is summarized.

2.1.1.1 Analysis of the First Order Model of the Cylindrical Dielectric Resonator Antenna (CDRA)

Resonant modes excited in dielectric resonators are divided into transverse electric (TE), transverse magnetic (TM), and hybrid electromagnetic (HEM) modes. Furthermore, understanding the radiation patterns depending on the desired resonant modes is necessary for antenna designs. Figure 2.2 shows the field distributions of some modes as an example.

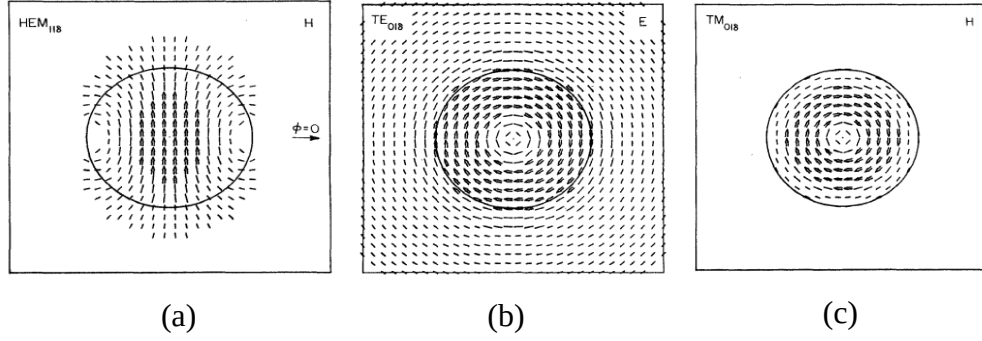


Figure 2.2 Field distributions of different modes of CDRA, (a) H field distribution of HEM_{118} mode (b) E field distribution of TE_{018} mode, and (c) H field distribution of TM_{018} mode © [1984] IEEE [40]

Here, by using Maxwell's equations and applying proper boundary conditions, we find the solutions of the Helmholtz equations, the components of electric and magnetic fields, in cylindrical coordinates. The dielectric constant should be high for the dielectric resonator. Therefore, the air-dielectric interface can almost resemble an open circuit, resulting in the creation of a resonator structure. The plane wave reflection coefficient is [41]:

$$\Gamma = \frac{\eta_0 - \eta}{\eta_0 + \eta} = \frac{\sqrt{\frac{\epsilon}{\epsilon_0}} - 1}{\sqrt{\frac{\epsilon}{\epsilon_0}} + 1} = \frac{\sqrt{\epsilon_r} - 1}{\sqrt{\epsilon_r} + 1} \cong 1 \quad (2.1)$$

where η , η_0 , ϵ , and ϵ_0 are the intrinsic impedance, intrinsic impedance of free space, and relative permittivity of DRAs, and permittivity of free space, respectively.

For the large values of dielectric permittivity, the reflection coefficient is almost +1. Therefore, the interface of the air-dielectric medium can be considered a perfect magnetic conductor (PMC) [41]. We use this approximation in solving the boundary condition problem. This is considered the first order-model of DR [41, 42]. The electric and magnetic field intensities are found by applying the boundary conditions when a DR is placed on a perfect electric conductor (PEC) plane, as shown in Fig 2.3.

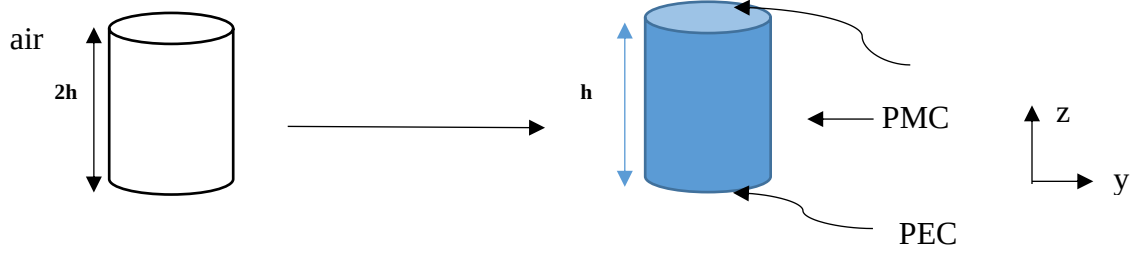


Figure 2.3 First order model of the dielectric resonator cavity

The vector potential functions A (magnetic vector potential) and F (electrical vector potential) using scalar wave equations are taken into consideration in order to do the modal analyses of TM and HEM modes and develop formulas for them. Here, the parameters of h , a , and β are the height of CDRA, the radius of CDRA, and the propagation constant, respectively. The boundary value problems are solved for the E and H fields using cylindrical coordinates. The terms ρ , ϕ , and z are the cylindrical coordinates.

TM^z Mode: $\nabla^2 A_z + \beta^2 A_z = 0 \quad \beta_\rho^2 + \beta_z^2 = \beta^2 \quad H_z = 0$

$$\left\{ \begin{array}{l} A = A_z(\rho, \phi, z) = A_{mn} J_m(\beta_\rho \rho) (c_4 \cos m\phi + d_4 \sin m\phi) (c_5 \cos \beta_z z + d_5 \sin \beta_z z) \\ F = 0 \end{array} \right. \quad (2.2)$$

$$E_\rho = \frac{A_{mn}}{j\omega\mu\epsilon} \beta_\rho J_m'(\beta_\rho \rho) (c_4 \cos m\phi + d_4 \sin m\phi) (d_5 \beta_z \cos \beta_z z - c_5 \beta_z \sin \beta_z z) \quad (2.3)$$

$$E_\phi = \frac{A_{mn}}{j\omega\mu\epsilon \rho} J_m(\beta_\rho \rho) (-c_4 m \sin m\phi + d_4 m \cos m\phi) (d_5 \beta_z \cos \beta_z z - c_5 \beta_z \sin \beta_z z) \quad (2.4)$$

$$E_z = \frac{A_{mn}}{j\omega\mu\epsilon} J_m(\beta_\rho \rho) (c_4 \cos m\phi + d_4 \sin m\phi) \times (c_5 (-\beta_z^2 + \beta^2) \cos \beta_z z + d_5 (-\beta_z^2 + \beta^2) \sin \beta_z z) \quad (2.5)$$

$$H_\rho = \frac{1}{\mu\rho} \frac{\partial A_z}{\partial \phi} = \frac{A_{mn}}{\mu\rho} J_m(\beta_\rho \rho) (-c_4 m \sin m\phi + d_4 m \cos m\phi) (c_5 \cos \beta_z z + d_5 \sin \beta_z z) \quad (2.6)$$

$$H_\varphi = \frac{-1}{\mu} \frac{\partial A_z}{\partial \rho} = -\frac{A_{mn}}{\mu} \beta_\rho J_m'(\beta_\rho \rho) (c_4 \cos m\varphi + d_4 \sin m\varphi) (c_5 \cos \beta_z z + d_5 \sin \beta_z z) \quad (2.7)$$

Boundary conditions:

DR is placed on PEC plane at $z=0$, so we have:

$$E_\varphi|_{z=0} = E_\rho|_{z=0} = 0 \longrightarrow d_5 = 0 \quad (2.8)$$

the air-dielectric medium is a PMC at $z=h$, so we have:

$$H_\varphi|_{z=h} = 0 \quad c_5 \cos \beta_z h = 0 \quad \beta_z h = \frac{(2l+1)\pi}{2} \quad l = 1, 2, \dots$$

for $l = 0$, we have $\beta_z h = \frac{\pi}{2}$, so E_z will be zero. Therefore, $l \neq 0$ (2-9)

$$H_\varphi|_{\rho=a} = 0 \longrightarrow J_m'(\beta_\rho a) = 0 \longrightarrow \beta_\rho = \frac{\chi'_{mn}}{a} \quad (2.10)$$

We have from dispersion equation:

$$\beta_\rho^2 + \beta_z^2 = \beta^2 \longrightarrow \left(\frac{\chi'_{mn}}{a}\right)^2 + \left(\frac{(2l+1)\pi}{2h}\right)^2 = \beta^2 \quad (2.11)$$

$$f = \frac{1}{2\pi\sqrt{\mu\epsilon}} \sqrt{\left(\frac{\chi'_{mn}}{a}\right)^2 + \left(\frac{(2l+1)\pi}{2h}\right)^2} \quad n = 1, 2, \dots, m = 0, 1, \dots, \text{ and } l = 1, 2, \dots \quad (2.12)$$

The electric and magnetic field components of TM^z modes when the DR placed on PEC plane are found:

$$E_\rho = \frac{A_{mn}}{j\omega\mu\varepsilon} \beta_\rho J_m' \left(\frac{\chi'_{mn}}{a} \rho \right) (c_4 \cos m\varphi + d_4 \sin m\varphi) \left(-c_5 \beta_z \sin \frac{(2l+1)\pi}{2h} z \right) \quad (2.13)$$

$$E_\varphi = \frac{A_{mn}}{j\omega\mu\varepsilon} J_m \left(\frac{\chi'_{mn}}{a} \rho \right) (-c_4 m \sin m\varphi + d_4 m \cos m\varphi) \left(-c_5 \beta_z \sin \frac{(2l+1)\pi}{2h} z \right) \quad (2.14)$$

$$E_z = \frac{A_{mn}}{j\omega\mu\varepsilon} J_m \left(\frac{\chi'_{mn}}{a} \rho \right) (c_4 \cos m\varphi + d_4 \sin m\varphi) \times \left(c_5 (-\beta_z^2 + \beta^2) \cos \frac{(2l+1)\pi}{2h} z \right) \quad (2.15)$$

$$H_\rho = \frac{A_{mn}}{\mu\rho} J_m \left(\frac{\chi'_{mn}}{a} \rho \right) (-c_4 m \sin m\varphi + d_4 m \cos m\varphi) \left(c_5 \cos \frac{(2l+1)\pi}{2h} z \right) \quad (2.16)$$

$$H_\varphi = -\frac{A_{mn}}{\mu} \beta_\rho J_m' \left(\frac{\chi'_{mn}}{a} \rho \right) (c_4 \cos m\varphi + d_4 \sin m\varphi) \left(c_5 \cos \frac{(2l+1)\pi}{2h} z \right) \quad (2.17)$$

$$H_z = 0 \quad (2.18)$$

Hybrid Mode: HEMmn

$$H_z \approx 0$$

$$\left\{ \begin{array}{l} A = A_z(\rho, \varphi, z) = A_{mn} J_m(\beta_\rho \rho) (c_6 \cos m\varphi + d_6 \sin m\varphi) (c_7 \cos \beta_z z + d_7 \sin \beta_z z) \quad (2.19) \\ F = F_z(\rho, \varphi, z) = F_{mn} J_m(\beta_\rho \rho) (c_8 \cos m\varphi + d_8 \sin m\varphi) (c_9 \cos \beta_z z + d_9 \sin \beta_z z) \quad (2.20) \end{array} \right.$$

$$\begin{aligned} E_\rho &= \frac{A_{mn}}{j\omega\mu\varepsilon} \beta_\rho J_m'(\beta_\rho \rho) (c_6 \cos m\varphi + d_6 \sin m\varphi) (d_7 \beta_z \cos \beta_z z - c_7 \beta_z \sin \beta_z z) \\ &+ \frac{-1}{\varepsilon\rho} F_{mn} J_m(\beta_\rho \rho) (-c_8 m \sin m\varphi + d_8 m \cos m\varphi) (c_9 \cos \beta_z z + d_9 \sin \beta_z z) \end{aligned} \quad (2.21)$$

$$E_\varphi = \frac{A_{mn}}{j\omega\mu\varepsilon\rho} J_m(\beta_\rho\rho)(-c_6m \sin m\varphi + d_6m \cos m\varphi)(d_7\beta_z \cos \beta_z z - c_7\beta_z \sin \beta_z z) + \frac{1}{\varepsilon} F_{mn}\beta_\rho J_m'(\beta_\rho\rho)(c_8 \cos m\varphi + d_8 \sin m\varphi)(c_9 \cos \beta_z z + d_9 \sin \beta_z z) \quad (2.22)$$

$$E_z = \frac{A_{mn}}{j\omega\mu\varepsilon} J_m(\beta_\rho\rho)(c_6 \cos m\varphi + d_6 \sin m\varphi)(c_7(-\beta_z^2 + \beta^2) \cos \beta_z z + d_7(-\beta_z^2 + \beta^2) \sin \beta_z z) \quad (2.23)$$

$$H_\rho = \frac{A_{mn}}{\mu\rho} J_m(\beta_\rho\rho)(-c_6m \sin m\varphi + d_6m \cos m\varphi)(c_7 \cos \beta_z z + d_7 \sin \beta_z z) + \frac{1}{j\omega\mu\varepsilon} F_{mn}\beta_\rho J_m'(\beta_\rho\rho)(c_8 \cos m\varphi + d_8 \sin m\varphi)(d_9\beta_z \cos \beta_z z - c_9\beta_z \sin \beta_z z) \quad (2.24)$$

$$H_\varphi = -\frac{A_{mn}}{\mu} \beta_\rho J_m'(\beta_\rho\rho)(c_6 \cos m\varphi + d_6 \sin m\varphi)(c_7 \cos \beta_z z + d_7 \sin \beta_z z) + \frac{1}{j\omega\mu\varepsilon\rho} F_{mn} J_m(\beta_\rho\rho)(-c_8m \sin m\varphi + d_8m \cos m\varphi)(d_9\beta_z \cos \beta_z z - c_9\beta_z \sin \beta_z z) \quad (2.25)$$

$$H_z = \frac{F_{mn}}{j\omega\mu\varepsilon} J_m(\beta_\rho\rho)(c_8 \cos m\varphi + d_8 \sin m\varphi)(c_9(-\beta_z^2 + \beta^2) \cos \beta_z z + d_9(-\beta_z^2 + \beta^2) \sin \beta_z z) \approx 0 \quad (2.26)$$

$$\longrightarrow F_{mn} \approx 0 \text{ very small value} \quad (2.27)$$

Boundary conditions:

The air-dielectric medium is a PMC at $\rho = a$, so we have:

$$H_\varphi|_{\rho=a} = 0, \quad J_m'(\beta_\rho a) = 0, \quad \beta_\rho = \frac{\chi'_{mn}}{a} \quad n = 1, 2, \dots \text{ and } m = 0, 1, \dots \quad (2.28)$$

DR is placed on PEC plane at $z=0$, so we have:

$$\longrightarrow E_\varphi|_{z=0} = 0 \quad d_7 = 0 \text{ and } c_9 = 0 \quad (2.29)$$

$$H_\varphi|_{z=h} = 0, \longrightarrow c_7 \cos \beta_z z = 0, \quad \beta_z h = \frac{(2l+1)\pi}{2} \quad l = 0, 1, 2, \dots \quad (2.30)$$

The Electric and magnetic fields of HE^z modes when the DRA placed on the PEC plane are:

$$\begin{aligned} E_\rho = & \frac{A_{mn}}{j\omega\mu\varepsilon} \beta_\rho J_m' \left(\frac{\chi'_{mn}}{a} \rho \right) (c_6 \cos m\varphi + d_6 \sin m\varphi) \left(-c_7 \beta_z \sin \frac{(2l+1)\pi}{2} z \right) + \\ & \frac{-1}{\varepsilon\rho} F_{mn} J_m \left(\frac{\chi'_{mn}}{a} \rho \right) (-c_8 m \sin m\varphi + d_8 m \cos m\varphi) \left(d_9 \sin \frac{(2l+1)\pi}{2} z \right) \end{aligned} \quad (2.31)$$

$$\begin{aligned} E_\varphi = & \frac{A_{mn}}{j\omega\mu\varepsilon\rho} J_m \left(\frac{\chi'_{mn}}{a} \rho \right) (-c_6 m \sin m\varphi + d_6 m \cos m\varphi) \left(-c_7 \beta_z \sin \frac{(2l+1)\pi}{2} z \right) + \\ & \frac{1}{\varepsilon} F_{mn} \beta_\rho J_m' \left(\frac{\chi'_{mn}}{a} \rho \right) (c_8 \cos m\varphi + d_8 \sin m\varphi) \left(d_9 \sin \frac{(2l+1)\pi}{2} z \right) \end{aligned} \quad (2.32)$$

$$E_z = \frac{A_{mn}}{j\omega\mu\varepsilon} J_m \left(\frac{\chi'_{mn}}{a} \rho \right) (c_6 \cos m\varphi + d_6 \sin m\varphi) \left(c_7 (-\beta_z^2 + \beta^2) \cos \frac{(2l+1)\pi}{2} z \right) \quad (2.33)$$

$$\begin{aligned} H_\rho = & \frac{A_{mn}}{\mu\rho} J_m \left(\frac{\chi'_{mn}}{a} \rho \right) (-c_6 m \sin m\varphi + d_6 m \cos m\varphi) \left(c_7 \cos \frac{(2l+1)\pi}{2} z \right) + \\ & \frac{1}{j\omega\mu\varepsilon} F_{mn} \beta_\rho J_m' \left(\frac{\chi'_{mn}}{a} \rho \right) (c_8 \cos m\varphi + d_8 \sin m\varphi) \left(d_9 \beta_z \cos \frac{(2l+1)\pi}{2} z \right) \end{aligned} \quad (2.34)$$

$$H_\phi = -\frac{A_{mn}}{\mu} \beta_\rho J_m' \left(\frac{\chi'_{mn}}{a} \rho \right) (c_6 \cos m\phi + d_6 \sin m\phi) \left(c_7 \cos \frac{(2l+1)\pi}{2} z \right) + \frac{1}{j\omega\mu\epsilon\rho} F_{mn} J_m \left(\frac{\chi'_{mn}}{a} \rho \right) (-c_8 m \sin m\phi + d_8 m \cos m\phi) \left(d_9 \beta_z \cos \frac{(2l+1)\pi}{2} z \right) \quad (2.35)$$

$$H_z = \frac{F_{mn}}{j\omega\mu\epsilon} J_m \left(\frac{\chi'_{mn}}{a} \rho \right) (c_8 \cos m\phi + d_8 \sin m\phi) (d_9 (-\beta_z^2 + \beta^2) \sin \beta_z z) \approx 0 \quad (2.36)$$

$$F_{mn} \approx 0 \quad (2.37)$$

2.1.1.2 Design Procedure of a Cylindrical DRA

Here, it is assumed that the CDRA is placed on the ground plane, as shown in Fig. 2.4. The parameters a , h , and ϵ_r are the radius, height, and relative dielectric constant of the cylindrical dielectric resonator antenna, respectively. Since the fields of CDRA do not have an exact solution, the equations of the resonance frequency for each excited mode have been derived based on numerical solutions and curve fitting [23].

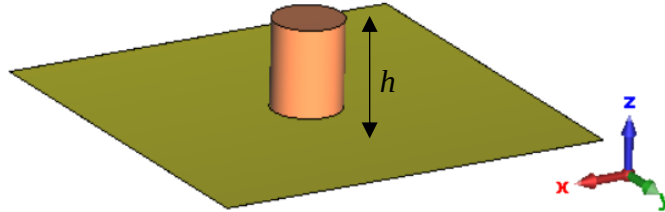


Figure 2.4 Geometry of the cylindrical DRA

The resonance frequency equations of three modes of CDRA, namely $HE_{11\delta}$, $TE_{01\delta}$, and $TM_{01\delta}$ can be found by Eq. (2.38-2.41) for discrete values of aspect ratio (a/h) [23, 43]:

$$k_0 a = \frac{6.324}{\sqrt{\epsilon_r + 1}} \left[0.27 + 0.36 \left(\frac{a}{2h} \right) + 0.02 \left(\frac{a}{2h} \right)^2 \right] \quad HE_{11\delta} \text{ mode} \quad (2.38)$$

$$k_0 a = \frac{2.327}{\sqrt{\varepsilon_r + 1}} \left[1 + 0.2123 \left(\frac{a}{h} \right) - 0.00898 \left(\frac{a}{h} \right)^2 \right] \quad TE_{01\delta} \text{ mode} \quad (2.39)$$

$$k_0 a = \frac{\sqrt{3.83^2 + \left(\frac{\pi a}{2h} \right)^2}}{\sqrt{\varepsilon_r + 2}} \quad TM_{01\delta} \text{ mode} \quad (2.40)$$

To design the CDRA, the proper permittivity and resonance frequency are assumed, and then the graph of $k_0 a$ is plotted using equations (2.38-2.40) for the selected mode. Similarly, the graph of $k_0 a$ is plotted using Eq. (2.41) [23] versus the aspect ratio (a/h) for some values of h_{cm} . Finally, the value of (a/h) and following that the radius of the DRA is established by where these lines intersect.

$$k_0 a = \frac{f_{GHz} h_{cm} \cdot (a/h)}{4.7713} \quad (2.41)$$

Here, the radius and aspect ratio of (a/h) are determined based on the given frequency of 64 GHz and the dielectric constant of 10 as an example. As can be seen in Fig. 2.5, the equations are plotted, and the intersection of these lines shows the proper aspect ratio. For instance, the intersection of the $h=0.4$ mm line and K_{0a} on graphs for $TE_{01\delta}$ mode shows the $(a/h) = 1.77$ and as a result, the acceptable radius is $a = 0.708$ mm.

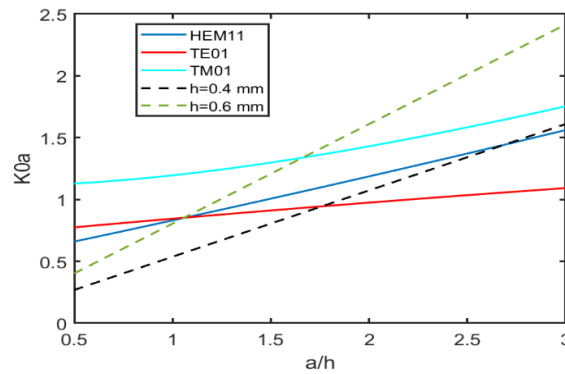


Figure 2.5 Calculated K_{0a} vs (a/h) for cylindrical DRA designs at 64 GHz for $\varepsilon_r = 10$

2.1.2 Rectangular Dielectric Resonator Antenna

In this section, the design procedure of a rectangular DRA is summarized.

The higher degree of design freedom of the rectangular dielectric resonator over cylindrical DRAs is one of its benefits, i.e., two degrees of freedom due to independently choosing the ratios (w/h) and (w/d) [23]. The geometry of the rectangular DRA is shown in Fig. 2.6.

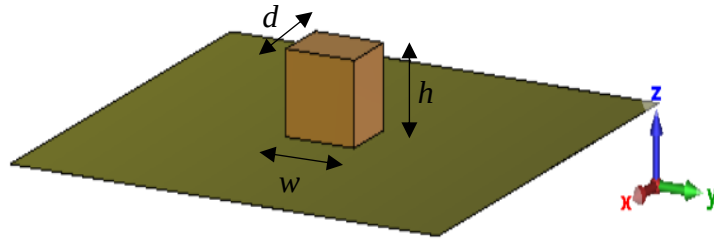


Figure 2.6 Geometry of the rectangular DRA

The parameters of w , h ($0.5b$), and d are the width, height, and depth of the rectangular dielectric resonator antenna, respectively. In Fig. 2.6, the excited TE modes of the rectangular DRA placed on the ground radiate like short magnetic dipoles [23].

A model of the dielectric waveguide is used to estimate the resonance frequency of the rectangular dielectric resonator. The following transcendental equation can be used to determine the resonance frequency of $TE_{\delta mn}$ [44]:

$$K_x \tan\left(\frac{k_x d}{2}\right) = \sqrt{-k_x^2 + k_{mn}^2 (\epsilon_r - 1)} \quad (2.42)$$

$$k_x^2 + k_y^2 + k_z^2 = k_{mn}^2 \quad (2.43)$$

$$k_z = \frac{n\pi}{b}, \quad k_y = \frac{m\pi}{w}, \quad k_{mn} = \frac{2\pi f_{mn}}{c} \quad (2.44)$$

c is the speed of light.

Figure 2.7 shows the simulated field distribution of the rectangular antenna with the dimensions of $w=d=5$ mm and $h=30$ mm for the $TE_{\delta 15}$ fed using a slot at the resonance frequency of 11 GHz.

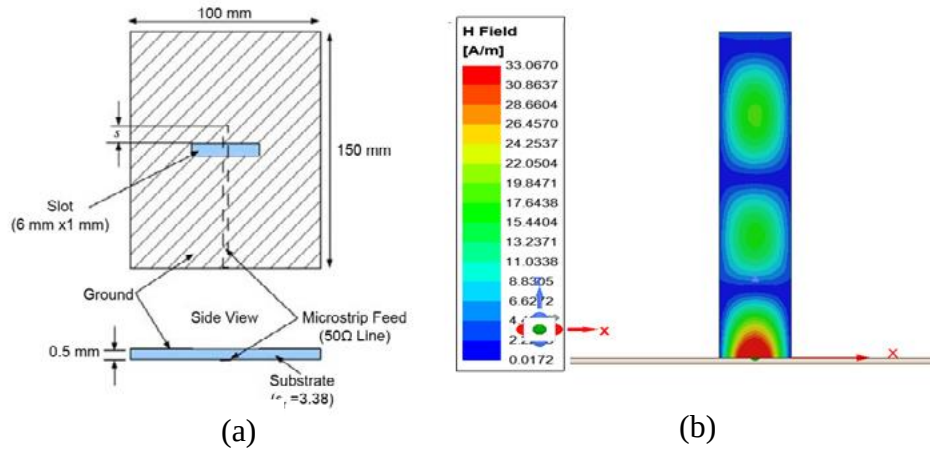


Figure 2.7 (a) Geometry of the microstrip-fed aperture © [2011] IEEE [44], and (b) Simulated H field distribution using HFSS tool v2021

2.2 Feeding Techniques of DRAs

As mentioned above, one of the advantages of the DRA is its capability to excite different modes using various feeding techniques. The feeding placement location where the field distributions are strong can be used to determine the excited mode [23]. Several feeding methods are reviewed in this section. Employing a probe-feed approach can cause the broadside radiation patterns shown in Fig. 2.8 (a) [45] and also create a unilateral radiation pattern [28]. Furthermore, a designed array including CDRA's fed by a single probe-fed microstrip patch has been reported in [46]. However, a direct probe feeding DRA can result in fabrication error and inappropriate impedance matching at high frequencies [26]. As can be seen in Fig. 2.8 (b), a microstrip line can be utilized to excite the DRA modes as a planar feeding network with the benefit of an easy fabrication process [47].

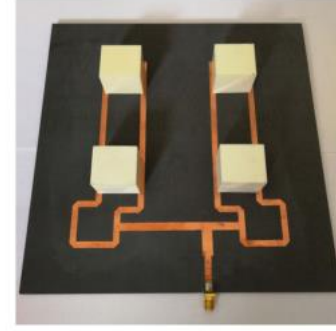
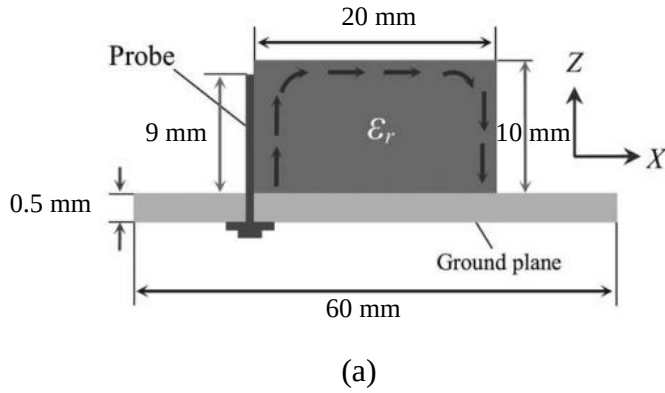


Figure 2.8 (a) $HEM_{11\delta}$ mode of a CDRA by the probe feeding technique © [2015] IEEE [45],
 (b) Rectangular DRA array prototype © [2018] IEEE [47]

Additionally, feeding the DRA using the slot cut on the ground plane, aperture coupling is a commonly reported technique [48, 49]. The CDRA designed using this technique is shown in Fig. 2.9. The antenna is simulated with the High Frequency Structure Simulator (HFSS) v2021 and CST Microwave Studio v2021 tools. The resonance frequency is 68 GHz, and a bandwidth of 4.8 GHz is achieved. The dimensions of the structure, DRA radius and height of a and h , slot widths of W_{slot} and W_{slot} , the thickness of the substrate h_s , and the used materials are mentioned in Table. 2.2.

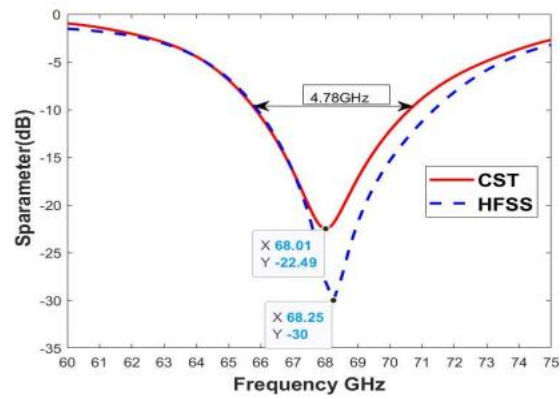
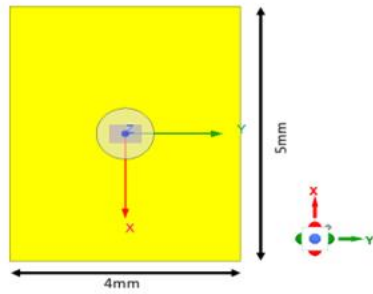


Figure 2.9 (a) Geometry of the CDRA, and (b) obtained S_{11} parameter

Table 2.2 The dimensions of optimized microstrip-fed aperture DRA parameters

Dimensions (DRA)	Dimensions (slot)	Dimensions(substrate)	Dimensions (MSL)
$a=0.5\text{mm}$	$W_{xslot}=0.35\text{mm}$	$h_s=0.254\text{ mm}$	Width=0.25 mm
$h=0.6\text{ mm}$	$W_{yslot}=0.55\text{mm}$	$W_x=5\text{mm}, W_y=4\text{ mm}$	Thickness=0.018
$\epsilon_r=10, \tan\delta=0.002$		$\epsilon_r=3, \tan\delta=0.001$	

Here, some feeding methods are mentioned, but these excitation systems have significant feeding losses in the millimeter-wave frequency range [50]. Substrate integrated waveguide (SIW) is a potential solution to the excitation of dielectric resonator antennas at high frequencies having the capabilities of integration with planar circuits and preventing radiation loss [51, 52]. In addition, the conduction loss brought on by the feeding network can be reduced with the aid of SIW technology [53]. Figure 2.10 shows two different antenna arrays fed by SIW technology. The structures shown in Fig. 2.10 (a) and (b) have an impedance bandwidth of 4.7% at the frequency range of 35.7–39.2 GHz [51] and that of 2.4% at the frequency range of 9.88–10.12 GHz [54]. In the next chapter, the antenna array designed using SIW technology will be presented.

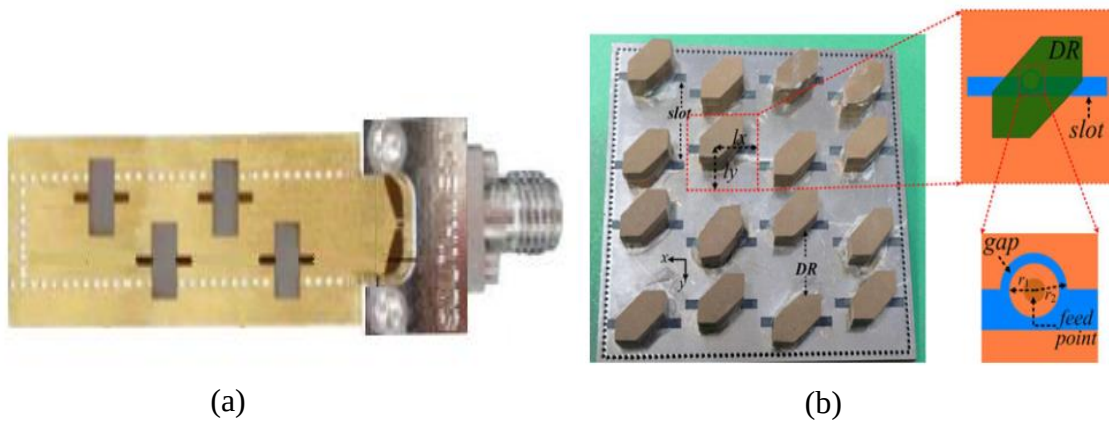


Figure 2.10 (a) Fabricated prototype series-fed SIW DRA © [2011] IEEE [51],
(b) Geometry of the DRA array © [2018] IEEE [54]

2.2.1 Array feed networks

A series-feed network and a corporate-feed network are two common array feed networks shown in Fig. 2.11 to excite DRAs as examples. In a parallel feed network, the network is symmetric and the line length for each element is equal, resulting in the frequency independence characteristic [55]. A series feed network is a simple construction with the potential to scan frequencies in an array configuration. The shortcoming of this feed network is the attenuation of the signal for the consecutive elements [55, 56].

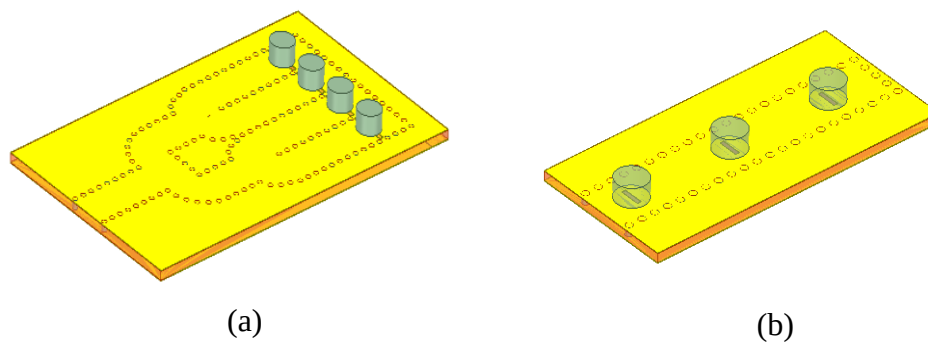


Figure 2.11 Array feeding networks. (a) Parallel feed and (b) Series feed

2.3 Bandwidth Enhancement Techniques

As mentioned in Table 2.1, one of the advantages of DRAs is their capability to achieve high bandwidth. Numerous different techniques have been reported to increase the DRAs bandwidth. In order to achieve a wide bandwidth, DRA is frequently constructed with different shapes, such as a rack shaped MIMO DRA (impedance bandwidth of 101.87%) at the frequency range of 54–10.89 GHz [57], truncated Tetrahedron DRA with a 40% wide bandwidth [58], and an H-shaped structure [59]. Employing an air region to reduce the Q-factor [60], various air regions to excite higher-order DR modes [61] are other techniques to improve bandwidth. In addition, two separate resonant modes of a substrate-integrated dielectric resonator antenna (SIDRA), shown in Fig. 2.12 (a), can increase the antenna bandwidth thanks to the dielectric-via loading [62]. Stacking the DRAs [63] and combining the multiple resonances [64, 65] were presented as bandwidth improvement techniques.

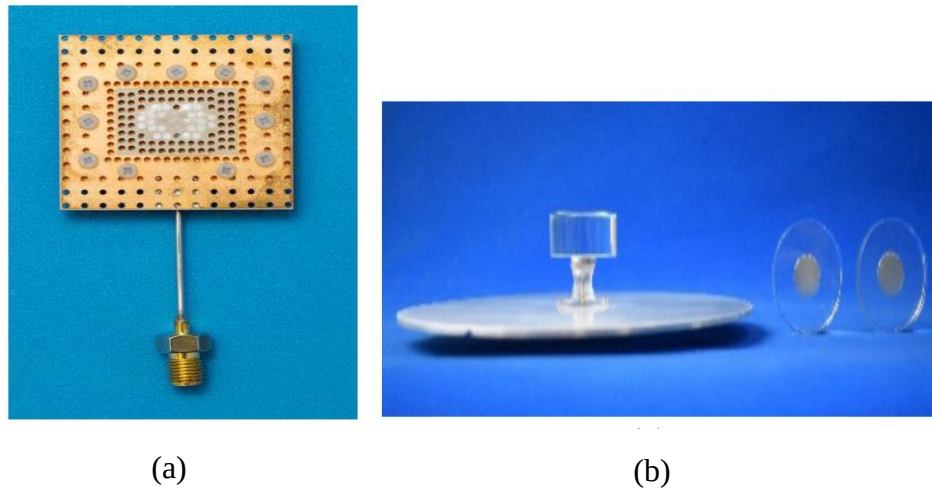


Figure 2.12 Photographs of prototypes (a) SIDRA © [2021] IEEE [62],
(b) Hybrid-glass DRAs [64]

Applying the perforated techniques to DRA designs was performed to achieve some advantages, including reducing the issues with aligning and gluing the separate DRA components [66, 67], especially in the array design, as well as improving the impedance bandwidth [68, 69]. Recently, employing wideband multi-rings DRA using the FDM 3-D printing approach have improved 10-dB impedance bandwidth to 60.2% at a frequency range of 4.3–8.0 GHz, as shown in Fig. 2.13 [70].

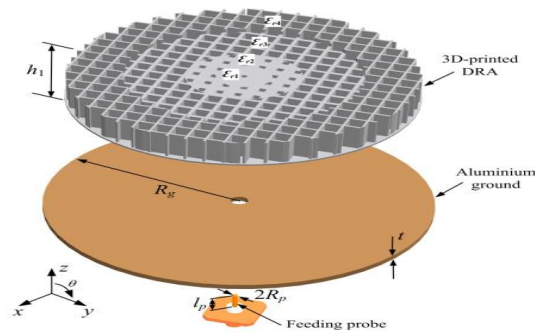


Figure 2.13 Configuration of the designed DRA using 3D printing technology
© [2019] IEEE [70]

2.4 Conclusion

In this chapter, a comprehensive review of previous works was presented. We reviewed the fundamental aspects of DRA and showed the design procedure for a cylindrical and rectangular DRA. Then the modal analysis of the cylindrical DRA was performed to develop formulas for them. The advantages and disadvantages of the DRA were explained and, finally, the feeding techniques for the excitation of the DRA modes were described.

CHAPTER 3 DIELECTRIC DRA DESIGN FED BY SUBSTRATE INTEGRATED WAVEGUIDE (SIW) AT MILLIMETER WAVE BAND

The unlicensed frequency range of 57 to 64 and 64 to 71 GHz in the millimeter-wave band is attractive to 5G applications [8, 71]. In wireless 5G applications, high gain antennas play a key role due to propagation loss and atmospheric absorption [10]. Array configurations [72] and the excitation of higher modes can be solutions for achieving high gain radiation [48]. In this chapter, a substrate integrated waveguide (SIW) considered as a feeding mechanism for the excitation of DRA modes is presented and designed. Then, a perforated cylindrical dielectric resonator antenna fed by the SIW feeding is introduced for millimeter wave applications. Subsequently, to design an array configuration using the parallel-fed antenna technique, a 6 dB power divider as a feeding mechanism for a SIW parallel-fed CDRA is designed, and lastly, the performance of the designed SIW parallel-fed cylindrical DRA is examined.

3.1 Design of the Substrate Integrated Waveguide as the Feeding Structure of DRA

In this section, a substrate integrated waveguide (SIW) for operating at the frequency range of 50 to 71 GHz is presented. SIW includes rows of metallic-vias drilled in the substrate with the permittivity of $\epsilon_r=3.24$, $\tan \delta = 0.0032$, and thickness of $h_s=0.239$ mm, which is covered by metal layers on top and bottom. The geometry of the SIW is shown in Fig. 3.1. If the spacing between metal vias is sufficiently small and radiation leakage can be ignored, the propagation properties of the SIW and its equivalent rectangular waveguide are similar [73].

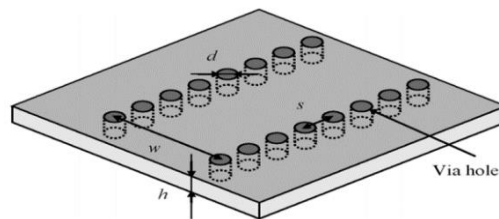


Figure 3.1 Geometry of the SIW © [2005] IEEE [74]

Here, the cut-off frequency of 42 GHz and the diameter of the via ($d=0.25$) mm are chosen. According to the design procedure that was presented in [74], we have :

Step1 :

$$s \leq 2d \longrightarrow s \leq 0.5 \text{ mm} \longrightarrow s = 0.48 \text{ mm} \quad (3.1)$$

Step 2 : we consider

$$d \leq w/5 \quad (3.2)$$

Step 3 : the width of the rectangular waveguide can be calculated by [41]

$$W_{eff} = \frac{c_0}{2f_c \sqrt{\epsilon_r}} = 1.98 \text{ mm} \quad (3.3)$$

Step 4 :

$$W_{eff} = w - 1.08 \frac{d^2}{s} + 0.1 \frac{d^2}{w} \longrightarrow w = 2.1 \text{ mm} \quad (3.4)$$

By following these steps, the via diameter, spacing between vias, and the width of SIW are $d = 0.25$ mm, $s = 0.48$ mm, and $W = 2.1$ mm, respectively.

Now, the optimization is done using the determined parameters. After optimization, the parameters of w , d , and s are 1.925 mm, 0.24 mm, and 0.481 mm, respectively. The geometry of the designed SIW and the magnitude of the electric field distribution at a sampling frequency of 66 GHz are shown in Fig. 3.2.

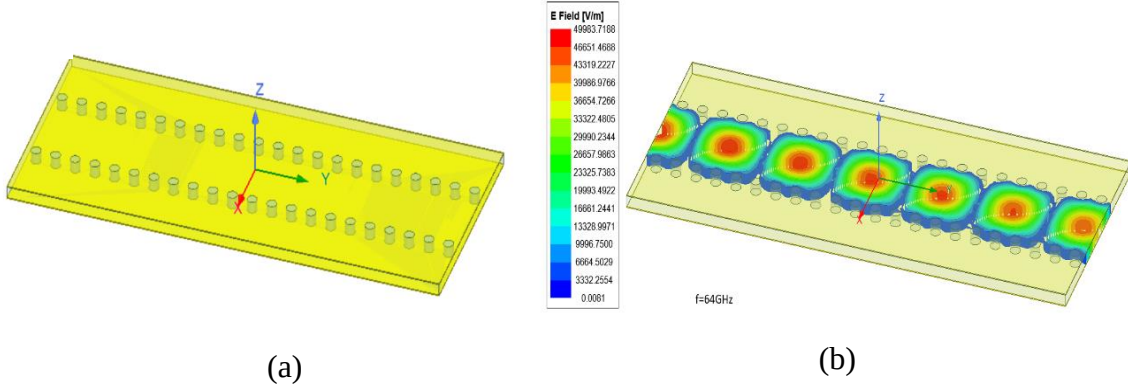


Figure 3.2 (a) Geometry of the designed SIW, (b) Magnitude of the electric field distribution at frequency of 64 GHz

To investigate the performance of the designed SIW, the s parameters (transmission and reflection coefficients) for the fundamental mode of TE_{10} are shown in Fig. 3.3. As can be seen, the values of S_{11} are almost less than -20 dB, and acceptable values of S_{21} (~ -0.74 dB) show a great transmission. In the next section, the designed SIW is used as a feeding mechanism of DRAs.

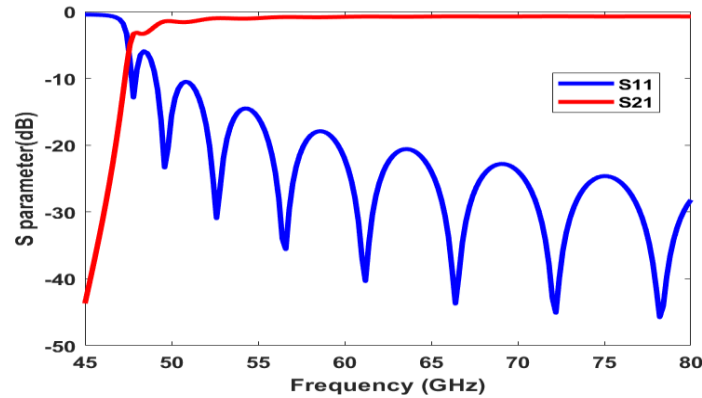


Figure 3.3 S parameters of the designed SIW

3.2 Design of SIW Fed-Cylindrical Dielectric Resonator Antenna

In this section, the designed SIW presented in the previous section is used as a feeding mechanism for DRAs. The energy is coupled into the resonator antenna through a slot [75] cut on the broad wall of the SIW and positioned in the center of the CDRA so that the HE_{11} mode of the CDRA is excited. The dimension parameters of slots are W_{xslot} and W_{yslot} . To couple the maximum power to

CDRA, the slot position of L_y is optimized. The values of $\lambda_g/4$ and $\lambda_g/2$ are selected for the transverse and longitudinal slot positions, respectively [50]. The substrate with a thickness of 0.239 mm is used in this design. The shown cylindrical dielectric antenna in Fig. 3.4(a) is made of the dielectric with the relative dielectric constant of $\epsilon_r=10$ and loss tangent of $\tan\delta=0.002$. The parameters a and h are the radius and height of CDRA, respectively. The SIW in this design is a single TE_{10} mode waveguide. To investigate the features of the CDRA, the effect of slot position L_y on the impedance matching is shown in Fig. 3.4 (b). As can be seen, changing the position of the slot influences the impedance matching. To maintain a decent level of S_{11} , $L_y=1.8$ mm is selected.

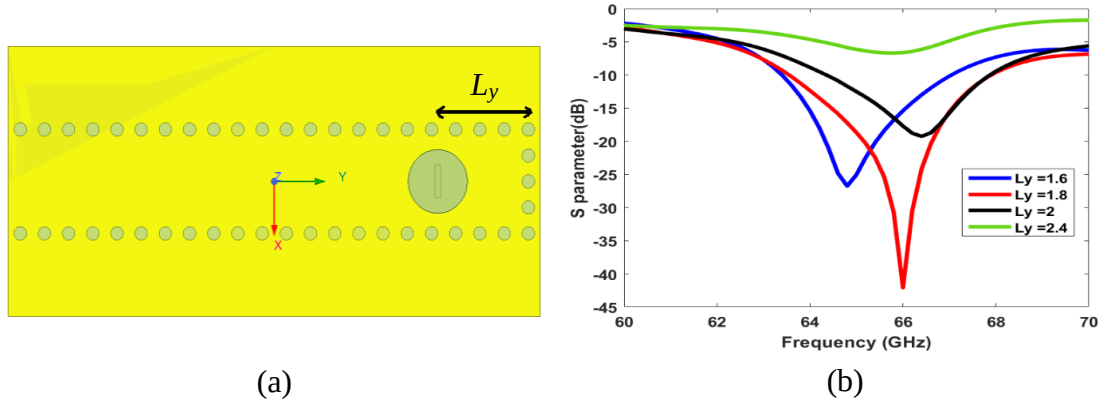


Figure 3.4 (a) SIW- based CDRA driven model, (b) The impact of slot position on the reflection coefficient

As mentioned in the previous chapter, changing the radius and height of DRA can shift the resonance frequency up or down, similar to a frequency tuning method. As a result, we change the height of DRA to find the resonance at 66 GHz. The obtained value of the DRA radius is 0.548 mm, as shown in Fig. 3.5 (a). The optimized parameters of the designed DRA can be found in Table 3.1. The radiation pattern of the designed antenna at two different planes, H (XZ-plane) and E (YZ-plane), at the frequency of 66 GHz is shown in Fig. 3.5 (b). The peak gain is 5.19 dB.

Table 3.1 The dimension size (in mm) of optimized SIW-DRA parameters

DRA	Slot	Substrate	SIW
$a=0.548$	$W_{xslot}=0.675$	$h_s=0.239$	$d_{via}=0.25$, $S=0.481$
$h=0.52$	$W_{yslot}=0.128$	$W_x=5$, $W_y=10.58$	Thickness=0.018

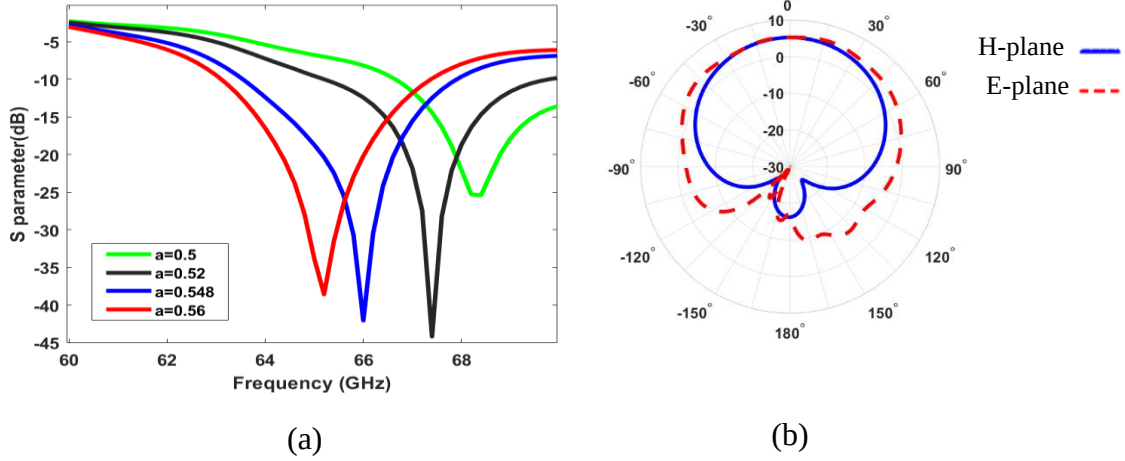


Figure 3.5 (a) The impact of DRA radius on the reflection coefficient, (b) radiation pattern of the CDRA at frequency of 66 GHz

3.3 Design of Perforated Dielectric Resonator Antenna

As mentioned in Chapter 1, the DRA is a potential candidate for wireless applications in the frequency range of 57-64 GHz in the millimeter wave band due to its advantages, namely high gain, low loss, high bandwidth, and fabrication capability in various shapes [23, 76, 77]. In addition, higher gain can be achieved by the excitation of higher order modes [44, 48, 78]. However, we experience difficulty in accurately locating and bonding DRAs, especially for the array design at high frequencies [23, 66, 67, 69]. Perforating the dielectric sheet can overcome this limitation and employing perforated techniques can also improve the impedance bandwidth in DRAs [68, 69], and reduce mutual coupling in THz antenna arrays [79], which means using these techniques can provide high gain and wideband antennas. To achieve this, a perforated dielectric resonator antenna is presented and shown in Fig. 3.6 The design procedure of the proposed structure is based on the design of a CDRA resonates at the frequency of 60 GHz and reducing the permittivity by creating a perforated region with the purpose of decreasing the quality factor and, consequently, increasing the bandwidth [23]. The DRA has a dielectric constant of $\epsilon_r=10.2$ and a loss tangent of $\tan \delta = 0.002$. By applying the tuning method, the parameters of the design structure are achieved. The lattice constant (a), thickness (t_1), diameter (d_1), and width (W_3) of the perforated DRA are 0.287 mm, 2.7 mm, 0.26 mm, and 2.8 mm, respectively. A SIW, which is a planar feeding mechanism, with the capability of integration with microstrip transmission lines is utilized [73]. The SIW substrate with the permittivity of $\epsilon_r=3.24$, $\tan \delta = 0.0032$, and thickness of $h_s = 0.239$ mm

is covered at the top and bottom by metal layers and has rows of metallic-vias drilled into it. The spacing between vias, the width of SIW, and the via diameter are $S = 0.5$ mm, $W_2 = 2.0$ mm, and $d_2 = 0.22$ mm, respectively. In order to couple the most energy to DRAs and excite the DRA mode, the perforated DRA is placed on the coupling slot that has been cut into the broad wall of the SIW, spaced $L_3 = 0.6$ mm from the centre of the SIW and $\lambda_g/4$ (1.05 mm) from short circuit vias [50].

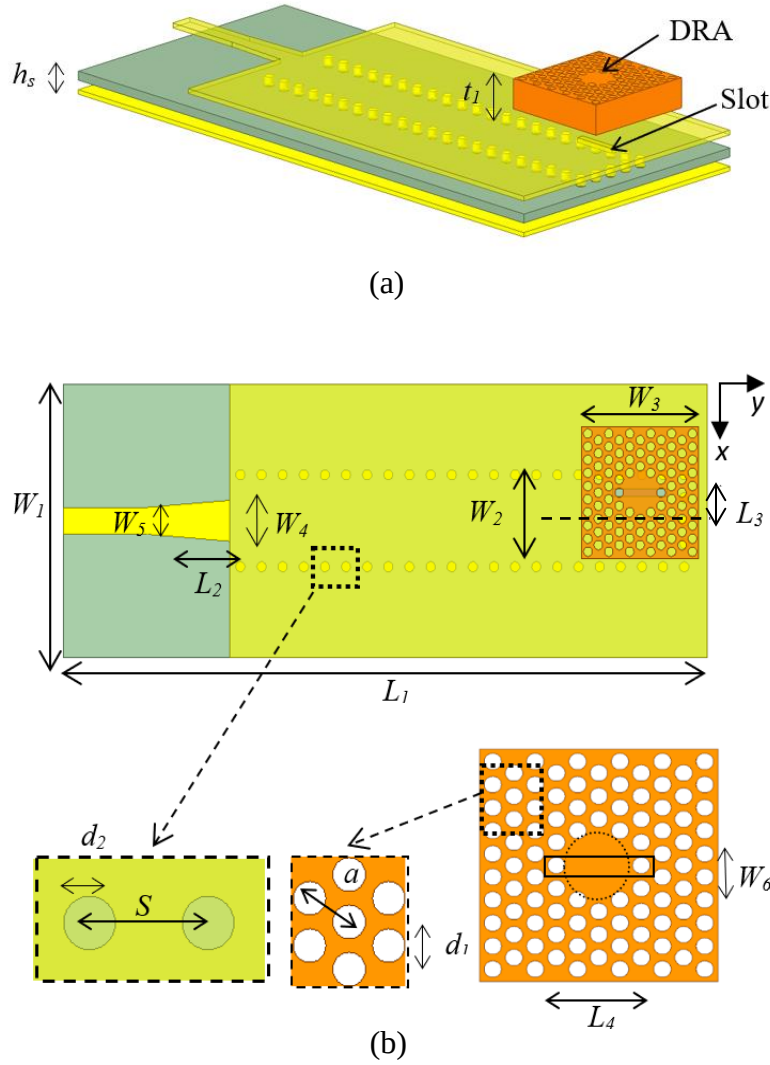


Figure 3.6 The geometry of the proposed structure (a) Isometric view (b) Top view

The slot dimension sizes are $W_6 = 0.18$ mm (width) and $L_4 = 1.2$ mm (length). The energy is transferred from the SMA connector to the SIW structure using a planar transition connected

between the SIW and microstrip line (MSL). The dimensions of the structure, namely the length of L_1 , the width of W_1 , the transition width of W_4 , and the microstrip width of W_5 are 15 mm, 6 mm, 0.75 mm, and 0.59 mm, respectively.

Here, the radiation characteristics of the perforated CDRA including the S-parameters, radiation pattern, and gain are studied using CST Microwave Studio software v2021. As can be seen in Fig. 3.7, the two main resonance frequencies are at 55 GHz and 60 GHz, thus covering completely the unlicensed spectrum of 57-64 GHz. The two resonance frequencies are merged together to make a wider bandwidth, i.e., 18.5% impedance bandwidth. The former resonance frequency is related to the slot, and the latter is for the perforated dielectric resonator antenna with an excited HEM_{113} mode. Fig. 3.8(a) shows the antenna gain in $\theta = 0^\circ$ direction (Z-axis). At a resonance frequency of 60 GHz, the peak gain is 8.6 dBi and an average of 8.2 dBi is obtained in the frequency range of 57-64 GHz. The peak gain is almost stable over this frequency range.

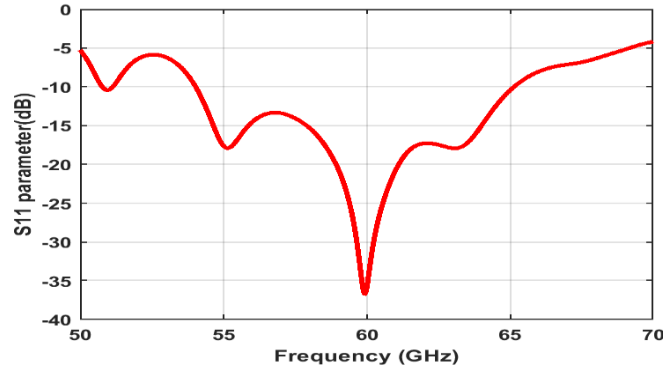


Figure 3.7 The simulated S_{11} parameter of the proposed perforated DRA

Figure 3.8(b) shows the radiation patterns (gain) of the presented perforated DRA at two distinct planes, E (XZ-plane) and H (YZ-plane), at the resonance frequency of 60 GHz. According to this figure, the radiation pattern is smooth and no ripple is observed. The H-plane radiation pattern is tilted due to the asymmetric structure, as expected.

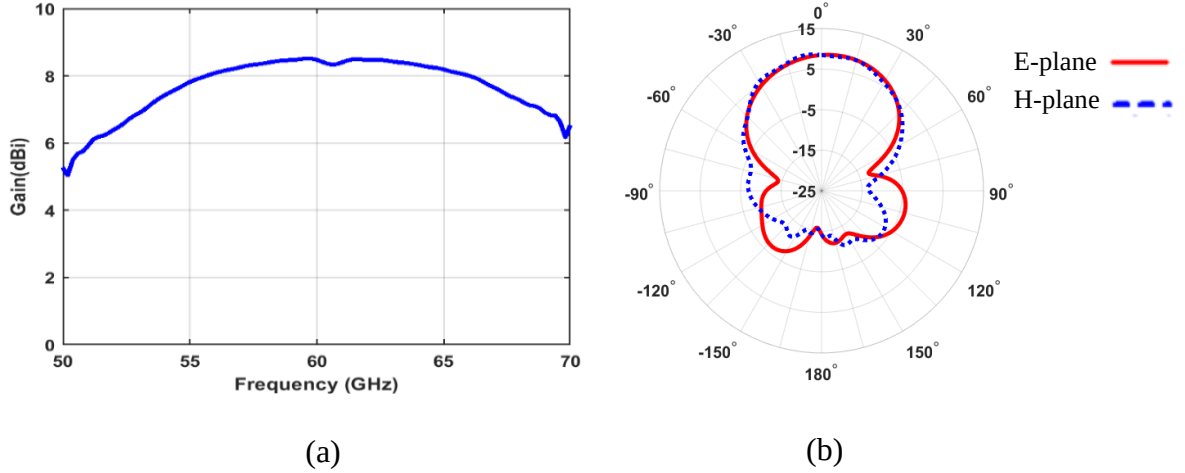


Figure 3.8 (a) The gain of the proposed antenna at the boresight direction ($\theta = 0^\circ$), and (b) The Radiation pattern (gain) of the proposed perforated DRA at the resonance frequency of 60 GHz

3.4 Design of 1×4 Substrate Integrated Waveguide -Power Divider

Employing a series feed of networks in large antenna array configurations is not appropriate due to their narrow bandwidth characteristic, although it can provide a lower degree of transmission loss in comparison with parallel feed networks [80]. Therefore, to design a parallel-fed, cylindrical dielectric resonator antenna, the need for a power splitter is inevitable. In this design, the power is split equally and with the same phases between the four CDRA's. Here, the SIW-power splitter parameters are the same as the designed SIW in Section 3.1. To achieve a minimum reflection, three posts are employed [81]. The structure of a 1×4 SIW-power splitter is shown in Fig. 3.9 (a). The dimension size (in mm) of parameters dy_1 , dy_2 , d_1 , d_2 , W_x , and W_y are 1.092, 1.23, 0.15, 0.18, 10, and 16.488, respectively. To study the performance of the designed power splitter, the s parameters are plotted in Fig. 3.10. The obtained results show great transmission ($S_{21} \sim -7.28$ dB) and an acceptable insertion loss. The -1.28 dB loss is related to the dielectric and SIW losses. For better clarification of the transmission, the electric field distribution at a frequency of 64 GHz is shown in Fig. 3.9(b). It confirms that the power is split equally.

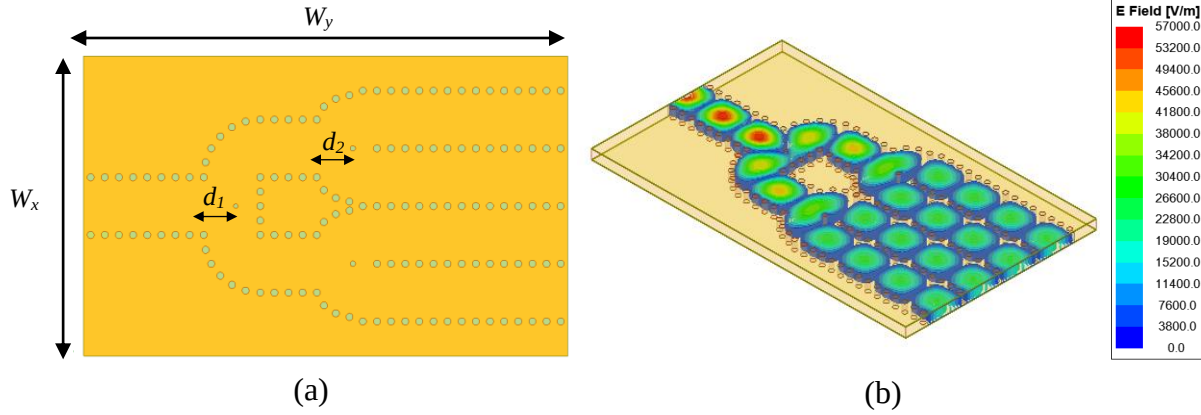


Figure 3.9 (a) Structure of a 1×4 (6 dB) SIW-power divider, (b)E-field distribution at a frequency of 64 GHz

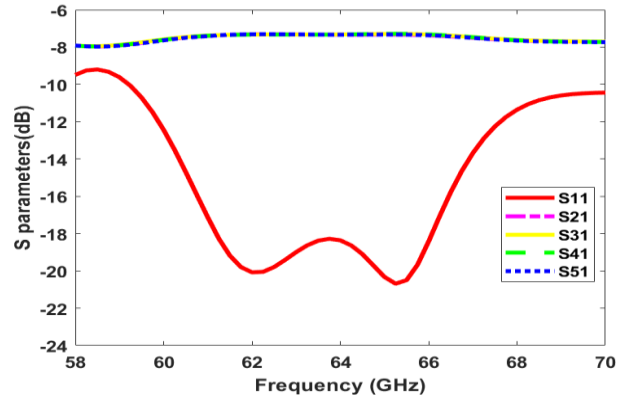


Figure 3.10 S parameters (dB) of the designed SIW based power divider

3.5 Design of SIW Parallel Fed-Cylindrical Dielectric Resonator Antennas Array

In recent years, the unlicensed spectrum of the millimeter wave band has attracted considerable attention. By considering the effects of path loss and atmospheric attenuation mentioned in Chapter 1, high-performance antennas with a proper gain are required. In this section, a linear array composed of four cylindrical DR antenna elements at the frequency range of 60-67 GHz is presented using the designed SIW-power splitter in the previous section, and then, the radiation performance of the array antenna is investigated.

To excite the CDRAs properly, longitudinal slots cut on the broad wall of the SIW are positioned at a distance of $\lambda_g/4$ that is almost 0.701 mm from short circuit vias to the slot position in order to couple the maximum energy to DRAs. The fundamental TE_{10} mode of the SIW-power splitter is

excited. All four DRA elements are excited equally and in the same phases. To achieve an appropriate performance, the parameters that influence the impedance matching such as slot width, slot length, and slot position are optimized. The optimized W_{xslot} and W_{yslot} of slots, the radius of CDRA, and its height are 0.78 mm, 0.189 mm, 0.56 mm, and 0.52 mm, respectively. As was mentioned before, changing the position of slots influences the impedance matching. Here, the slot position of L_y is 1.85mm. The proposed array configuration is shown in Fig. 3.11. Investigation of the radiation characteristics of the parallel-fed CDRA's array is performed using High Frequency Structure Simulator (HFSS) software v2021. According to the reflection coefficient S_{11} (dB) illustrated in Fig. 3.12(a), -10 dB bandwidth of the array configuration is 5.5 GHz (18%). The radiation pattern (gain) of the designed DRA array at resonance frequencies of 63.75 GHz is 9.19 dB at $\theta=0^\circ$ plane, as can be seen in Fig. 3.12(b).

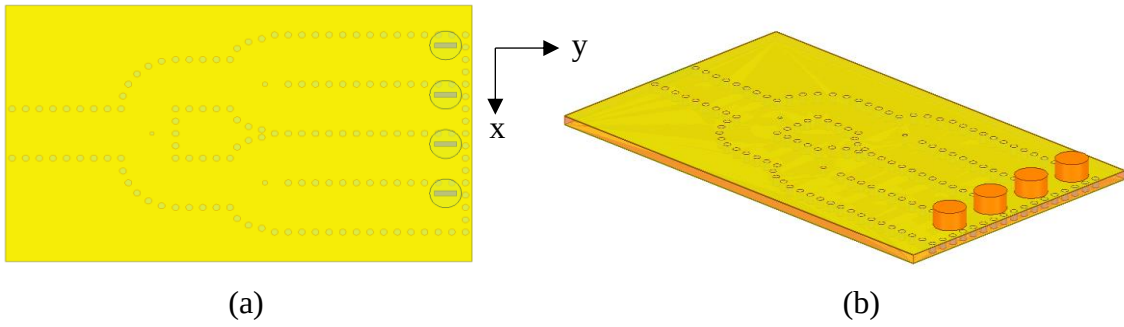


Figure 3.11 Structure of the SIW Parallel-Fed CDRA's (a) Top view (b) Isometric view

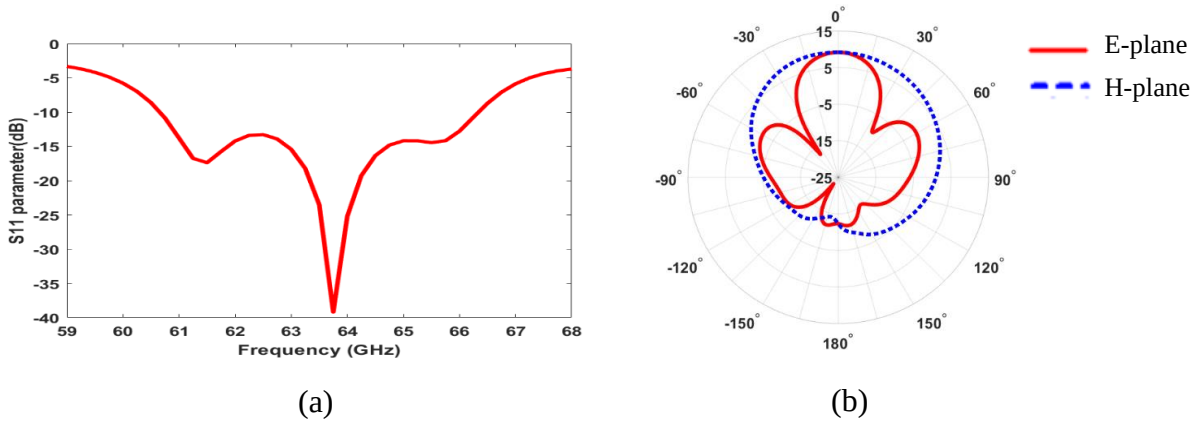


Figure 3.12 (a) Simulated reflection coefficient of parallel array configurations for the longitudinal slot position, (b) radiation pattern of the CDRA at a resonance frequency of 63.75 GHz

In the last part of this section, the performance of array antennas is examined when the MSL transition is added to the structure to transfer the power from the connector to the SIW structure. The geometry of the designed SIW parallel-fed CDRA with MSL transition is shown in Fig. 3.13(a). The reflection coefficient in Fig. 3.13 (b) shows a 6 GHz bandwidth (less than -10 dB), and two main resonance frequencies at 61.85 GHz and 64 GHz. Fig. 3.14 shows the radiation pattern (gain) of the designed DRA array with MSL transition at resonance frequencies of 61.85 and 64 GHz at two different planes, E (XZ-plane) and H (YZ-plane).

As can be seen, there are no considerable ripples in the radiation pattern after adding the MSL transition. However, the H-plane is tilted due to the asymmetric structure at the YZ-plane. The peak gains in the $\theta = 0^\circ$ direction are 8.8 dB and 8.95 dB at resonance frequencies of 61.85 and 64 GHz, respectively.

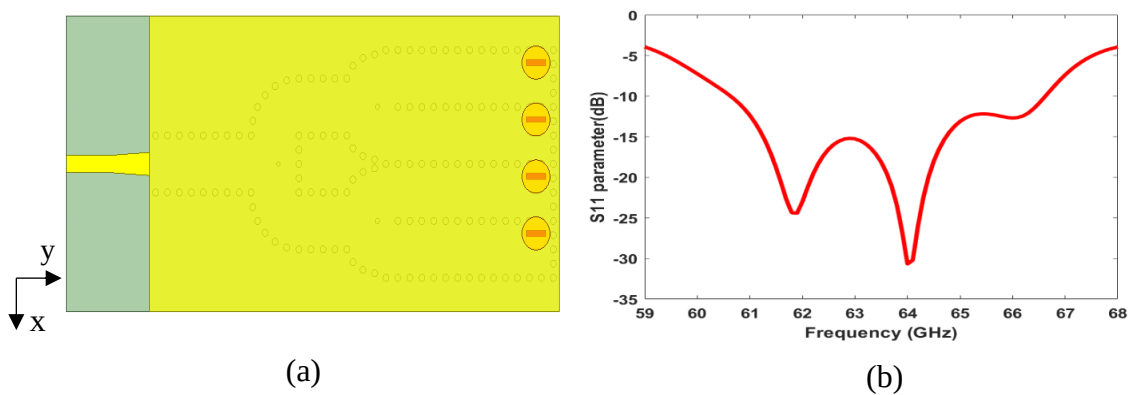


Figure 3.13 (a) Structure of SIW parallel fed-CDRAs with transition, (b) The simulated reflection coefficient of parallel fed CDRA's array with MSL transition

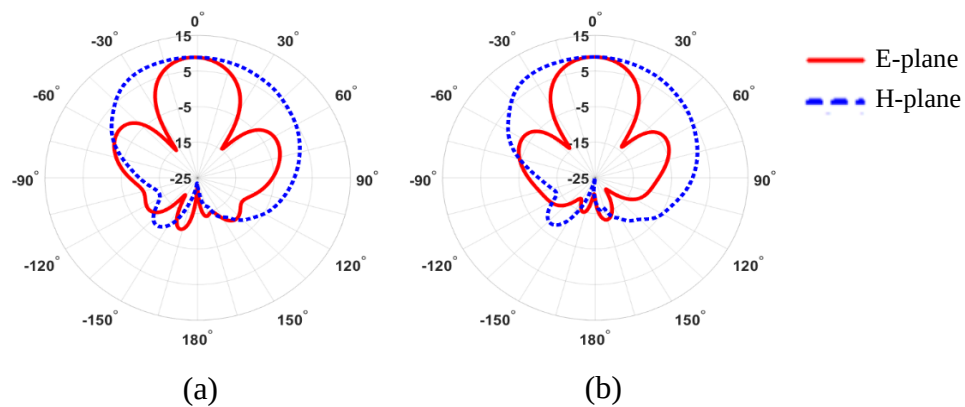


Figure 3.14 Radiation patter (gain) of parallel fed array with MSL transition at resonance frequencies of (a) 61.8 GHz and (b) 64 GHz

3.6 Conclusion

In this chapter, a substrate integrated waveguide (SIW) considered as a feeding mechanism for the excitation of cylindrical DRA modes was studied and designed. A high gain perforated dielectric resonator antenna fed by the SIW was then presented for millimeter wave applications. The designed antenna covers the unlicensed spectrum of 57-64 GHz with an 18.5% impedance bandwidth, and it shows an average gain of 8.2 dBi over the frequency range of 57-64 GHz. Next, to design an array configuration using the parallel-fed antenna technique, a 6 dB power divider as a symmetric feeding mechanism of the same phases at the output ports was used, and finally, the performance of the designed SIW parallel-fed cylindrical DRA was examined.

CHAPTER 4 DESIGN OF NOVEL DIELECTRIC BASED ANTENNAS INTEGRATED WITH LOW LOSS FEEDING SUBSTRATE INTEGRATED GUIDES

4.1 Introduction

In recent years, growing wireless data traffic and an increasing number of new applications have made the importance of Sub-THz and THz communications paramount as a viable technology in 6G and beyond to achieve high data rates [21, 82, 83]. However, the path loss and atmosphere attenuation are high, which significantly limits wireless communications at long distances, and as a result, high gain antennas are needed [19, 21, 84-86]. In this regard, various antenna designs based on different fabrication technologies have been reported. For instance, horn and reflector-based antennas show great performance at THz frequencies, but they are not suitable due to their large size [21]. The use of defective photonic crystal substrates in THz photoconductive antenna arrays can result in higher directivity and less mutual coupling [79]. At high frequencies, the printed circuit board (PCB) technology is fairly constrained due to large conductor losses and fabrication accuracy [87, 88], e.g., when substrates are laminated [72]. Additionally, substrate integrated waveguide (SIW) technology cannot be practical for frequencies higher than 200 GHz due to the fabrication complexity and the cost resulting from metal via diameters smaller than 50 μm [89]. Planar antennas with integration capability to other circuits are employed at Sub-THz frequencies [21, 90]. On chip antennas based on the complementary metal-oxide semiconductor (CMOS) technology gain attention, but they suffer from poor radiation efficiency and low gain [84, 91, 92]. Dielectric resonator antennas are an attractive solution for overcoming these disadvantages to some extent due to high gain and low loss advantages [23, 36, 93]. In addition, planar dielectric waveguides as a feeding method for the millimeter wave and THz integrated circuits can be utilized in antenna designs [94, 95]. The designed antenna using the planar dielectric waveguide feeding are reviewed in the next section, and based on that, the feeding mechanism for dielectric based antennas is proposed.

4.2 Designed Antennas Fed by Planar Dielectric Waveguides

Here, several antenna designs fed by planar dielectric waveguides are reviewed. A designed, millimeter wave, tapered antenna based on silicon-on-glass technology is shown in Fig. 4.1. The

antenna shows an average gain of 13.1 dB over the frequency range of 50-75 GHz [96]. Fig. 4.2 shows a dielectric rod antenna fed by a substrate integrated image guide (SIIG) at a frequency of 94 GHz with a measured gain of 12.8 dBi, but it was complex, and the E-plane pattern is tilted [97]. It is worth mentioning that tapered antennas are fragile at higher frequencies.

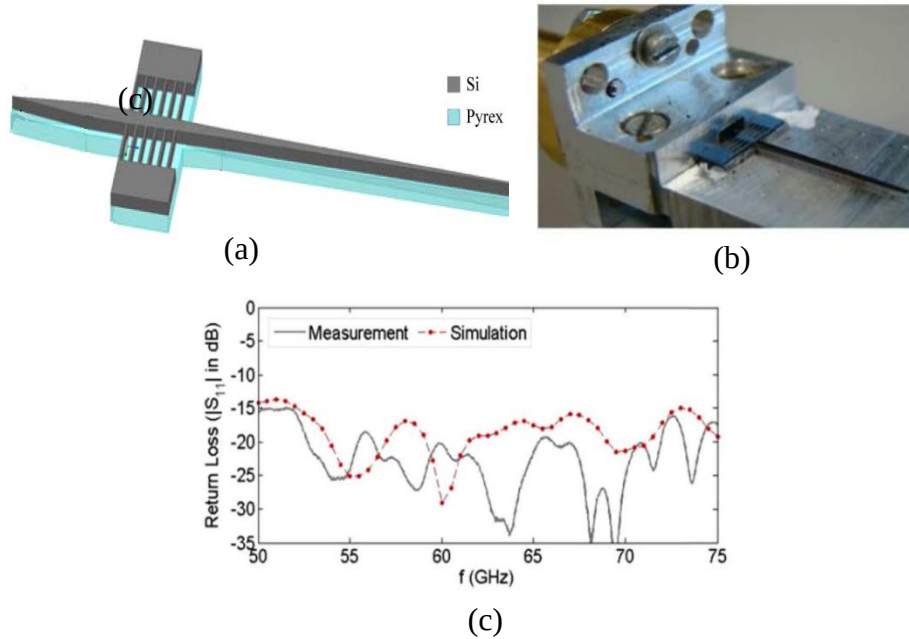


Figure 4.1 (a) Tapered SOG antenna, (b) Fabricated sample, and (c) Measured and simulated S_{11} parameters © [2014] IEEE [96]

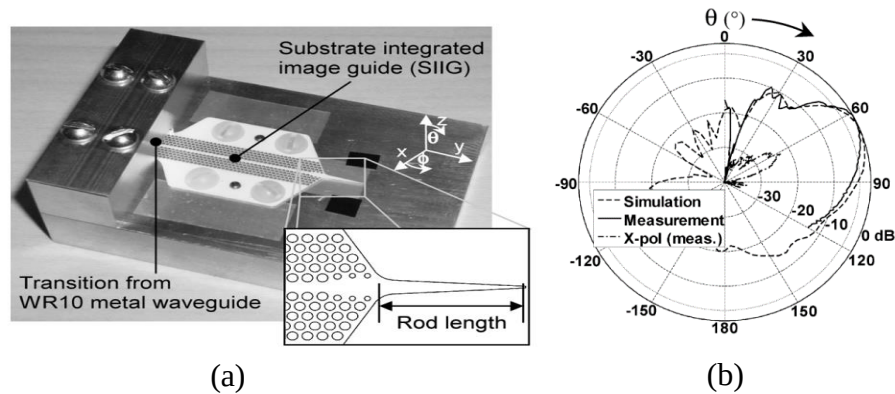


Figure 4.2 (a) Fabricated SIIG rod antenna at 94 GHz (b) E-plane radiation pattern in $\phi = 90^\circ$ © [2006] IEEE [97]

Dielectric image guides (DIGs) are also reported as a feeding mechanism of antennas. The geometry and S parameter of a 7-element DRA array fed by DIG are shown in Figs. 4.3 (a) and

(b), respectively. The antenna gain is 7.62 dBi at 10 GHz [98]. In addition, the array configuration and S_{11} parameter of a 7-element DRA array with a back reflector coupled to slot apertures fed by DIG are shown in Figs. 4.3 (c) and (d), respectively. The measured gain of 10.27 dBi at 10 GHz has been reported in [99].

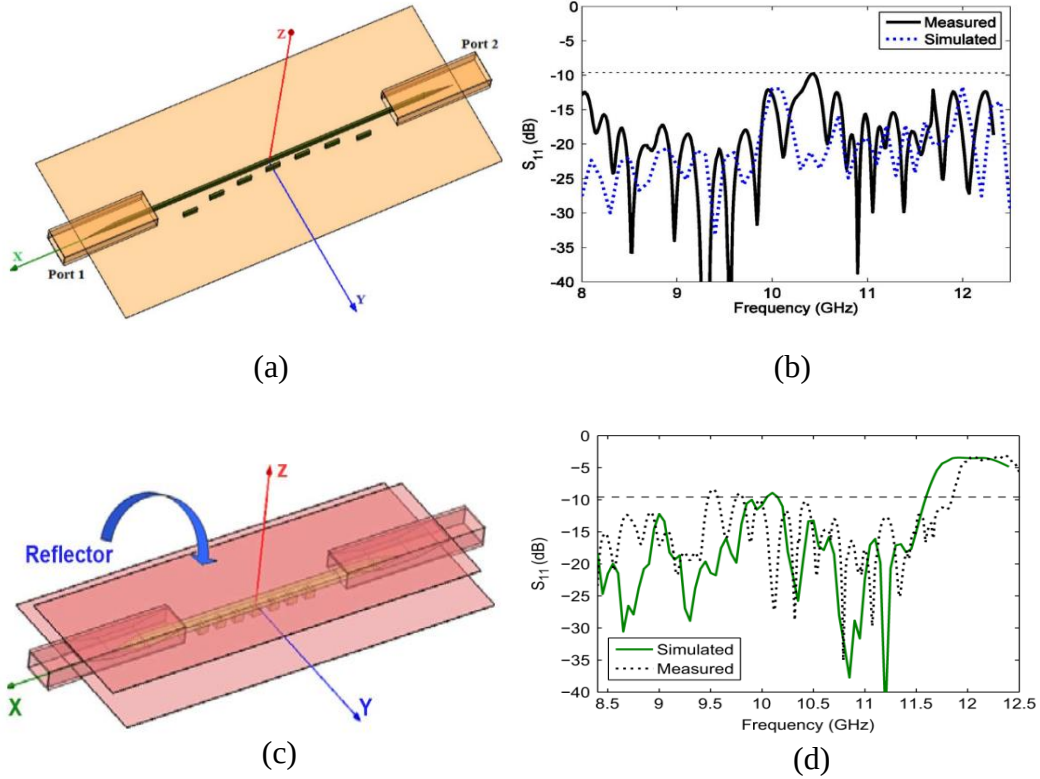


Figure 4.3 (a) Geometry of the 7-element DRA fed by DIG (b) Simulated and measured S_{11} parameter © [2010] IEEE [98], (c) Geometry of the 7-element DRA coupled to DIG through aperture slots with back reflector, (d) Simulated and measured S_{11} parameter © [2009] IEEE [99]

Recently, endfire radiation has been reported using the integration of a photonic crystal waveguide and dielectric resonator antenna from 311 to 331 GHz with a measured gain peak of 10.6 dBi, but there were deficiencies in its ground plane and supporting structure, as can be seen in Fig. 4.4. In this design, to improve impedance bandwidth to 6%, a matching air hole was employed [100]. To stabilize the designed antenna, a supporting structure based on silicon-on-glass technology has been proposed in [101]. Fig. 4.5 shows the geometry and performance of a DRA based on topological photonic crystals that do not need matching air holes. The impedance bandwidth and maximum gain are 3.46% and 7.4 dBi, respectively [102].

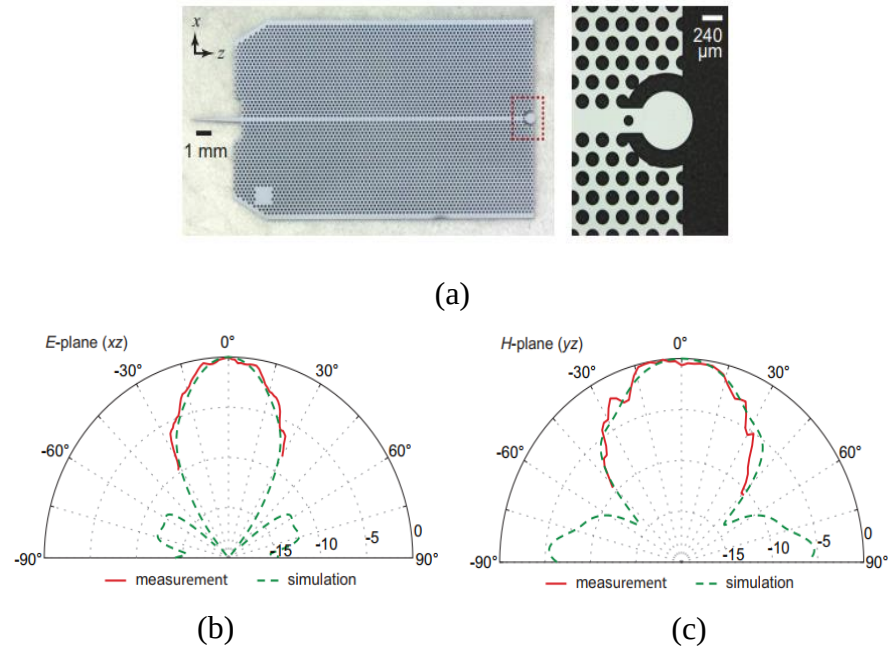


Figure 4.4 (a) Fabricated structure from a single silicon wafer, (b) and (c) Simulated and measured radiation patterns at 325 GHz in the xz plane and yz plane, respectively [100]

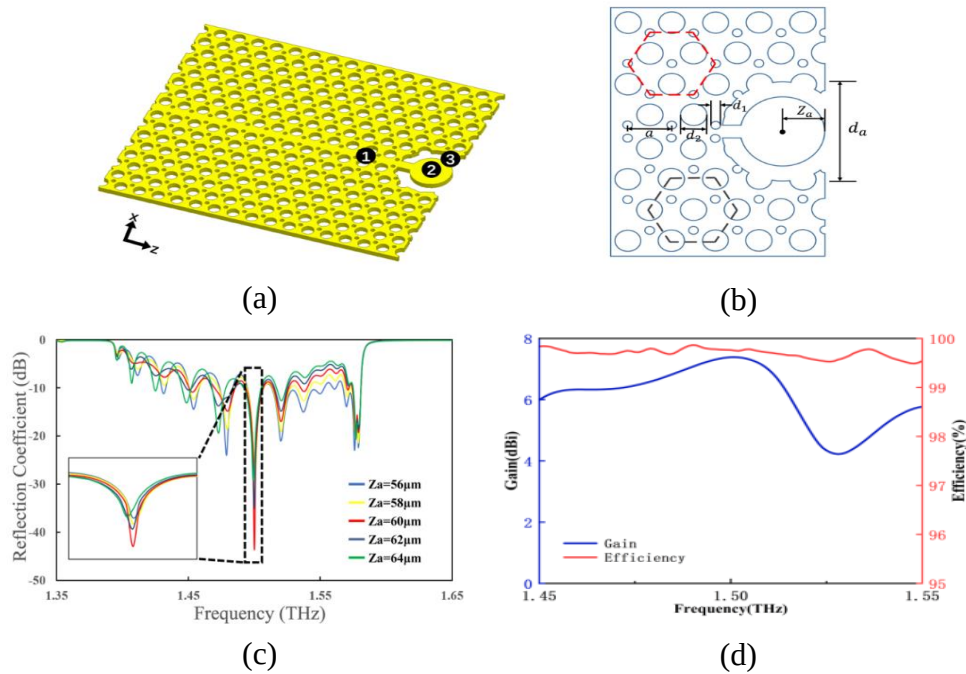


Figure 4.5 (a) Geometry of the DRA based on topological photonic crystals, (b) Top view, (c) Reflection coefficient, and (d) Gain and radiation efficiency, with the permission of AIP Publishing [102]

Despite attempts to design the dielectric antenna in the millimeter wave, Sub-THz, and THz bands, the new antenna design must have capabilities of integration with other devices, a planar form, and a low manufacturing cost. According to the presented literature review, we introduced a low loss feeding mechanism using a dielectric waveguide, and then we presented a novel dielectric base antenna integrated with the proposed feeding dielectric waveguide. Finally, the measurement and simulation results of the designed structures will be discussed.

4.3 Design of a Feeding Substrate-Integrated Guide in the Sub-THz Band

In this section, two different substrate integrated waveguides are presented at the frequency ranges of 140-180 GHz and 220-280 GHz. The first one is designed based on silicon-on-glass (SOG) technology [94]. The second one is designed based on laser micromachining, and will be employed as a feeding mechanism of the dielectric based antenna in the next section.

4.3.1 Design of a Substrate-Integrated Silicon Waveguide Supported by a Pyrex Layer at an Operating Frequency Range of 140-180 GHz

In this section, a feeding substrate integrated guide is proposed that could be a great candidate for coupling the energy to dielectric resonator antennas in the Sub-THz frequency band. The proposed waveguide consists of a perforated silicon waveguide supported by a Pyrex layer coated with a copper layer. To decrease loss, there is an air gap between the layers [94]. The proposed waveguide is shown in Fig. 4.6. WR6 rectangular waveguides are employed to couple the waves to the silicon waveguide through a tapered transition. The relative permittivity of the silicon and pyrex materials are 11.7 and 4.4, respectively. The radius, lattice constant, length, and width of the waveguides are $r=0.168$ mm, $a=0.513$ mm, $L=8.67$ mm, and $W=4.6$ mm, respectively. To investigate the performance of the proposed waveguide, the reflection and transmission of the proposed waveguide are investigated. The S parameters of the optimized waveguide are plotted in Fig. 4.7 at the frequency range of 130GHz to 200 GHz based on the conductivity of HR-Si for different resistivity values reported in [103]. The obtained results show the insertion loss varies in the range of 0.33 to 0.5 dB, and 0.8 to 1.05 dB for HR-silicon conductivity of 0.01 and 0.12 (s/m), respectively, at the frequency range of 140-180 GHz. In addition, the values of S_{21} parameter are

less than -20 dB at the frequency range of 140-180 GHz. According to the achieved results, the proposed silicon-based waveguide can be used for the Sub-THz applications in integrated systems.

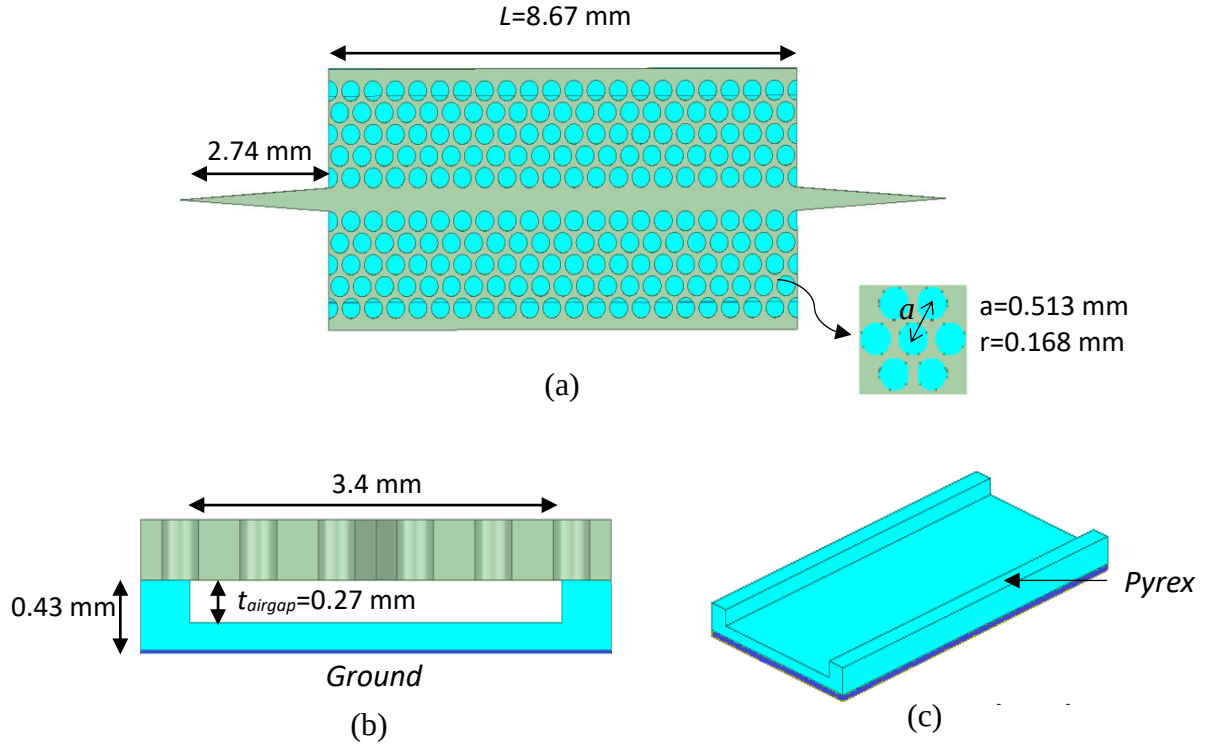


Figure 4.6 The 3D view of the proposed substrate integrated waveguide (a) Top view, (b) Side view, and (c) Perspective view

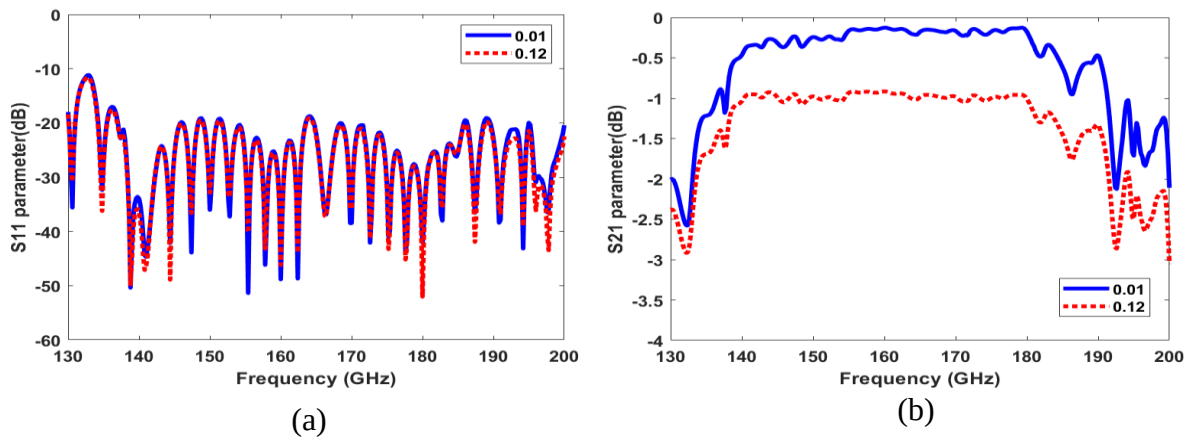


Figure 4.7 (a) S_{11} parameter and (b) S_{21} parameter (in dB) of the proposed substrate integrated waveguide for different values of HR-silicon conductivity

For better clarification of the proposed waveguide performance, the electric field distribution at the sampling frequency of 160 GHz is shown in Fig. 4.8. As can be seen, the magnitude of the electric field distribution is confined in the guiding channel and propagates, thereby confirming the performance of the waveguide.

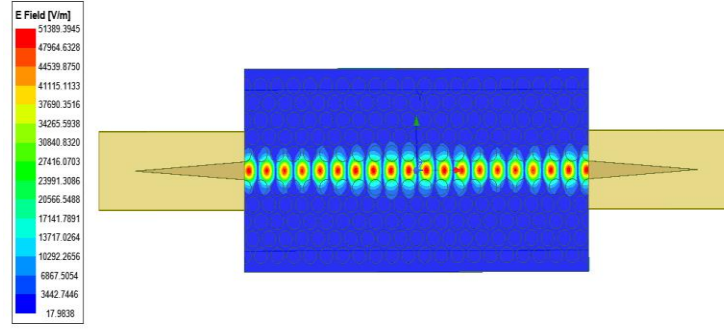


Figure 4.8 Magnitude of the electric field distribution at 160 GHz

In the presented structure, there is an air gap at the bottom of the waveguide, and according to the S-parameter plots for various gap thicknesses shown in Fig. 4.9, we can infer that we should consider a minimum thickness to obtain an acceptable transmission and reflection coefficient. The proposed waveguide is a basic design and will be used for a design of a Palanr dielectric waveguide to feed a dielectric-based antenna at a frequency of 250 GHz.

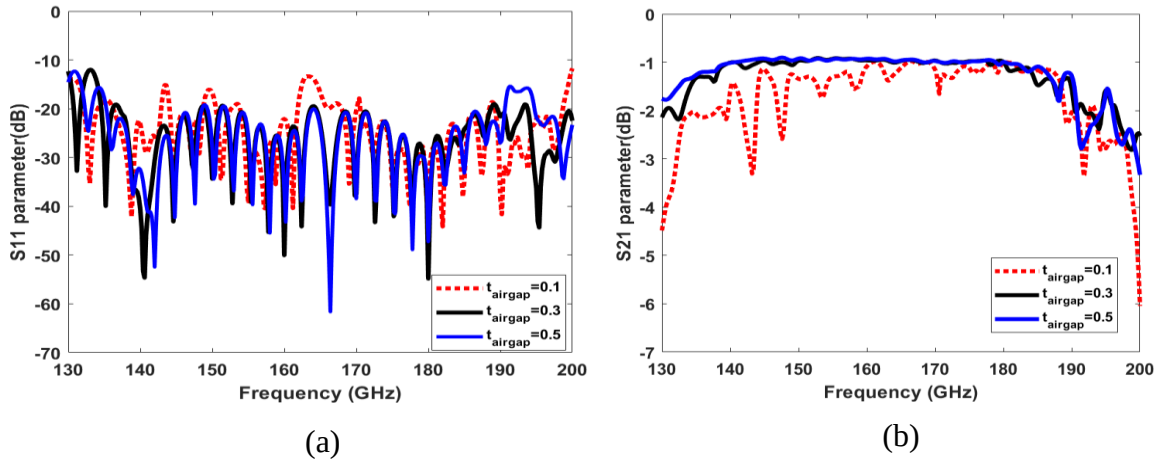
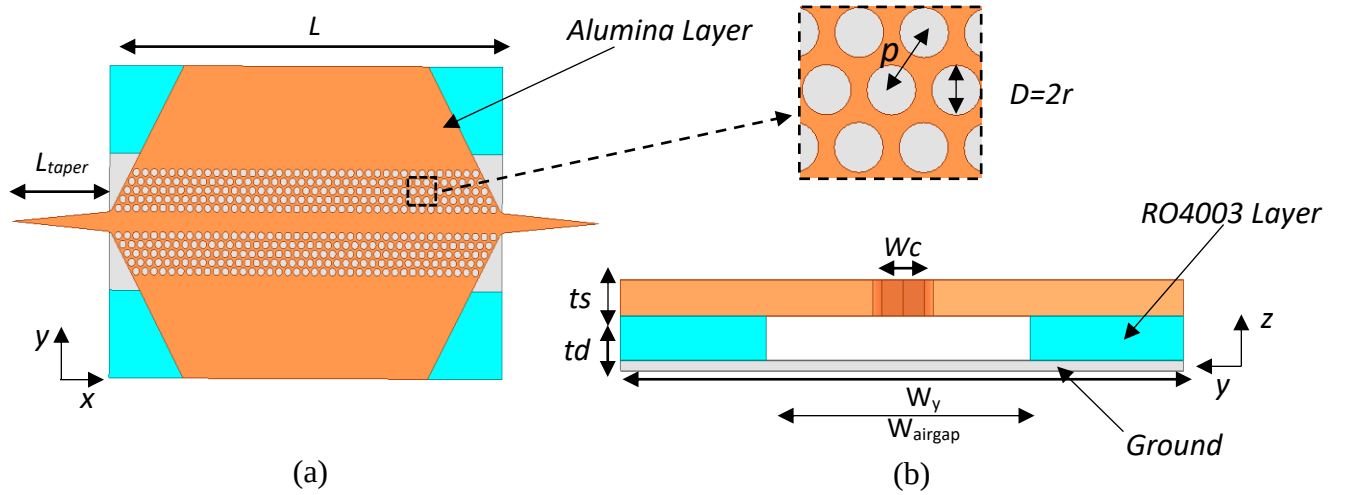


Figure 4.9 (a) S_{11} parameter and (b) S_{21} parameter (in dB) of the proposed substrate integrated waveguide for different air gap thicknesses

4.3.2 Design of a Substrate-Integrated Alumina Waveguide Supported at an Operating Frequency Range of 220-280 GHz

To couple the power to the dielectric based antenna and excite it, we presented a substrate-integrated alumina waveguide that operates at the frequency range of 230-280 GHz. The proposed waveguide structure consists of an alumina guiding channel and two symmetric air hole perforated regions on two sides of the guiding channel like SIIG [87, 104] . The alumina substrate with a relative permittivity of 9.8 is supported by a metal coated RO4003 layer with a permittivity of 3.55 and $\tan\delta=0.0027$. The high permittivity of the alumina layer results in the field confinement inside the guiding channel. Furthermore, there is an air gap between the guiding channel and the ground plane. The proposed structure is depicted in Fig. 4.10.



4.1.1.1 Simulation and Discussion

In this section, the proposed substrate-integrated guiding is studied, and the effect of geometrical and substrate parameters, namely the width of the guiding channel W_c , thickness of the alumina layer t_s , and thickness of the air gap t_{airgap} on the transmission coefficient is investigated. Then, the final design as a feeding mechanism of the DBA is presented based on the fabrication limitations and obtained simulation results.

Here, the parametric study is performed using a CST microwave studio v2021. The waveguide length, taper length, air gap width, air gap thickness, and width of the waveguides are $L=11$ mm, $L_{taper}=2.86$ mm, $t_{airgap}=0.305$ mm, $W_{airgap}=3.5$ mm and $W_y=10$ mm, respectively. The width of the air gap needs to be wider than the width of the guiding channel so that it does not increase the loss. The radius of $r=0.1$ mm, and lattice constant of $p=0.258$ mm are considered based on the approximated permittivity value of 5. Figure 4.11(a) shows the simulated reflection and transmission coefficients of the designed dielectric waveguide for different guiding channel widths of $w_c=0.2, 0.25$, and 0.3 mm. As can be seen from this figure, the frequency range of appropriate values of the transmission coefficient shifts downward with an increase in w_c . The S_{11} parameter for the value of $w_c=0.25$ mm is also less than -15 dB over the frequency range of 225-275 GHz. In addition, a similar trend is observed when the waveguide thickness t_s increases from 0.2 to 0.3 mm.

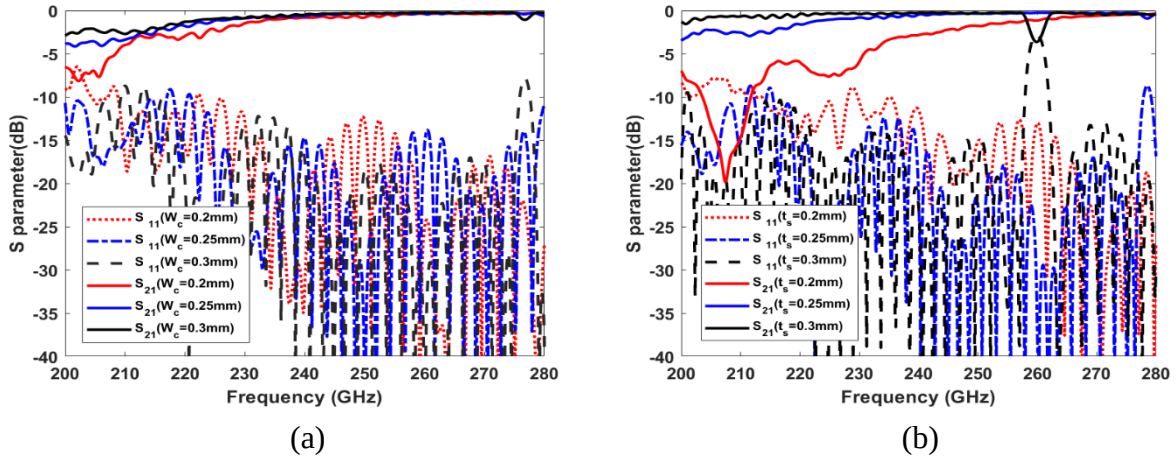


Figure 4.11 (a) Simulated reflection and transmission coefficients of the proposed waveguide as a function of frequency for different values of (a) guiding channel width (w_c) when the waveguide thickness is $t_s=0.25$ mm and (b) waveguide thickness (t_s) when the guiding channel width is $w_c=0.3$ mm

Fig. 4.12 (a) shows the simulated S-parameters of the waveguide for different air gap thicknesses of $t_{airgap}=0.025, 0.1$, and 0.3 mm. With reference to Fig. 4.12, the air gap thickness should be increased so that the proper transmission can be achieved. Moreover, another factor determined the thickness of the air gap is the DBA performance, including the side lobes and peak gain. Finally, the width of the air gap is studied from the S-parameter perspective. As can be seen in Fig. 4.12

(b), the air gap width underneath the guiding channel should be wide enough not to have an influence on the waveguide performance. The figure indicates that the S-parameters for values larger than 3 mm are almost the same.

Consequently, the guiding channel width of $w_c=0.25$ mm, air gap width of $w_c=3.5$ mm, and the waveguide thickness of $t_s=0.254$ mm are chosen based on the simulation results and manufacturing limit at Poly-Grames Research Center for the final design at the frequency range of 220-280 GHz.

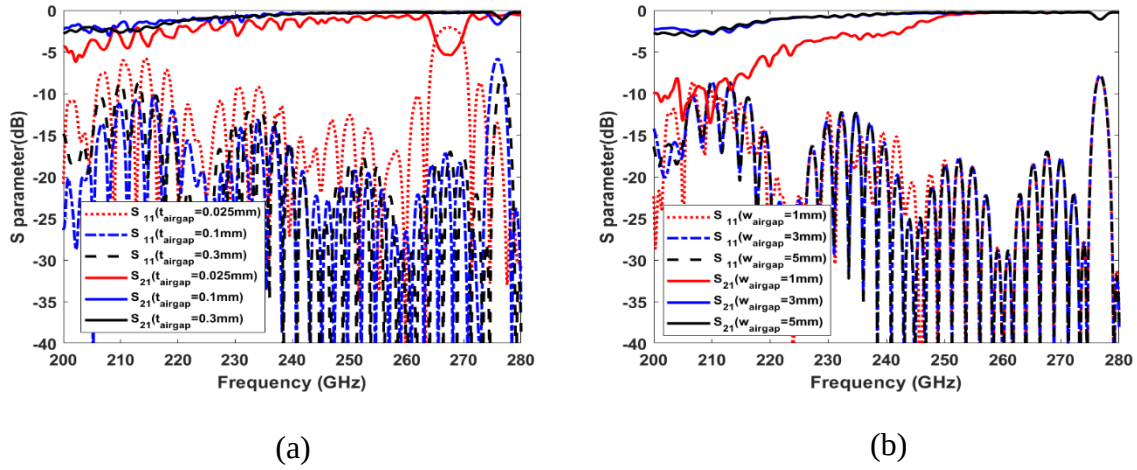


Figure 4.12 (a) Simulated reflection and transmission coefficients of the proposed waveguide as a function of frequency for different values of (a) air gap thickness (t_{airgap}) and (b) air gap width (w_{airgap}) when the guiding channel width and waveguide thickness are $t_s=0.25$ mm and $w_c=0.3$ mm

4.1.1.2 Fabrication Design

The final design of the proposed substrate integrated guide is shown in Fig. 4.13. The final design is considered based on the fabrication limitations of the laser drilling system and proper dimension sizes for the feeding mechanism of the DBA. The radius and lattice constant of the air holes perforation are $r=0.1$ mm and a lattice constant of $p=0.265$ mm. The air holes perforation is performed by laser micromachining. The optimized width of the guiding channel W_c , the thickness of the alumina layer t_s , and the height of the air gap t_d are 0.258 mm, 0.254 mm, and 0.305 mm, respectively. Additionally, a new metal fixture is designed and fabricated using the CNC machining technology to connect the designed waveguide to extenders of the measurement setup. Metal screws are employed to make the designed waveguide stable and avoid air gaps between layers.

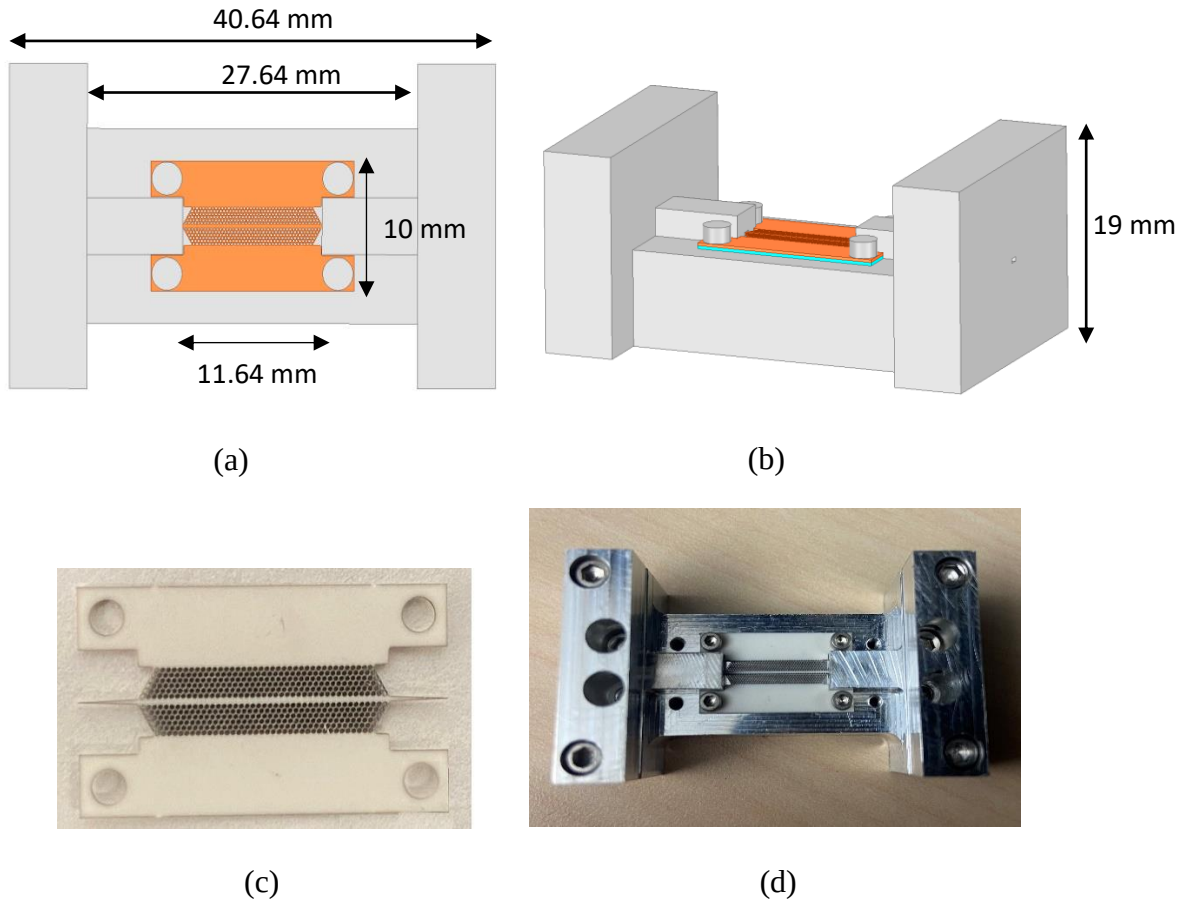


Figure 4.13 The designed short waveguide for the fabrication (a) Top view (b) Diametric view (c) Photograph of the fabricated Alumina waveguide, and (d) Photograph of the presented back to back DBA architecture

4.1.1.3 Measurement Results and Discussion

In this section, the S-parameters of the proposed waveguide are measured and compared with the obtained simulation results. The measurement system setup is shown in Fig. 4.14. The designed waveguide is connected to VDI extenders of the measurement setup through the designed fixture. A Vector Network Analyzer (VNA) equipped with VDI frequency extenders is used to measure S-parameters at the frequency range of 220-280 GHz, and then, the calibration is performed. Here, two proposed waveguides with the same dimensions but with different lengths are considered. The lengths of the short waveguide and long waveguide are 11.64 mm and 23.28 mm, respectively. The obtained simulated and measured results of the designed waveguides are illustrated in Fig. 4.15. As can be seen in this figure, the obtained results are symmetric, and there is good agreement

between measured and simulated values. The measured insertion loss of the short prototype varies between 2.27 dB and 3.84 dB over 240-280 GHz. Similarly, the measured insertion loss of the long prototype varies between 3.05 dB and 5.53 dB over 240-280 GHz. Additionally, with reference to the figure, the proper reflection coefficient is achieved for the short waveguide. The different insertion losses of simulation and measurement results can be attributed to the lateral misalignment, unwanted air gap, and fabrication errors.

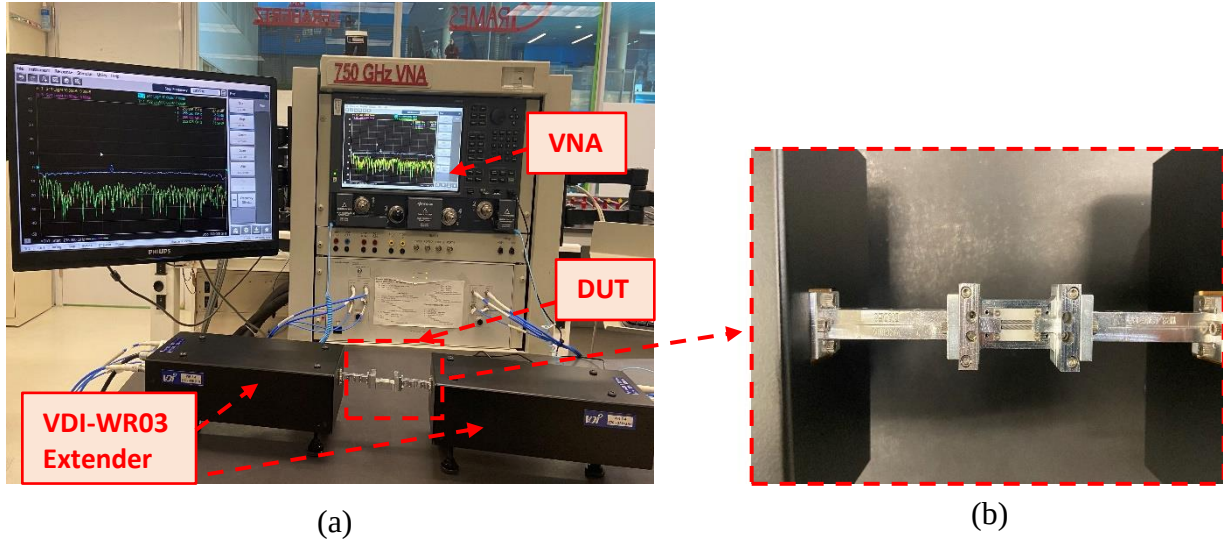


Figure 4.14 (a) THz measurement set-up (b) the proposed waveguide under test (DUT)

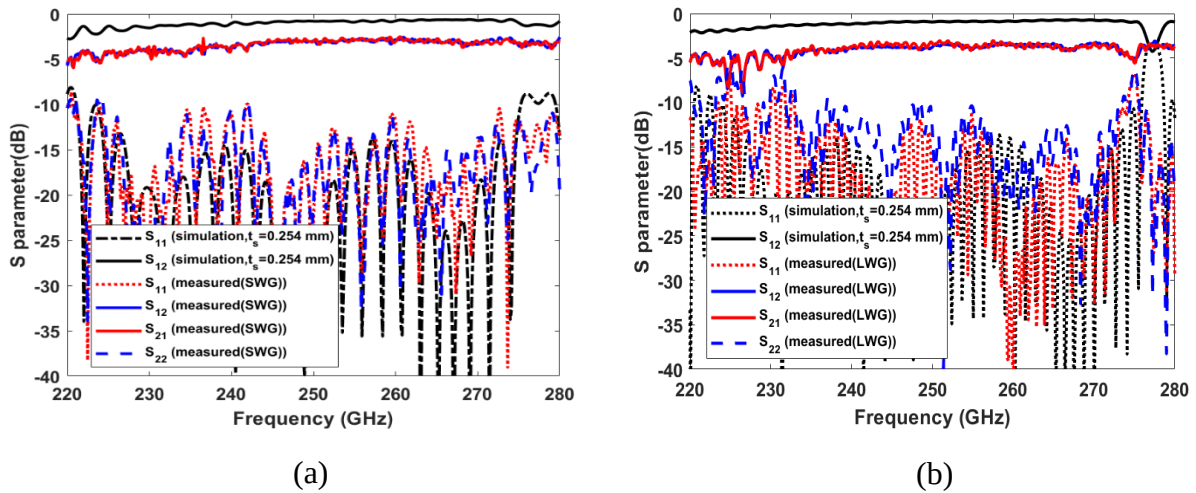


Figure 4.15 The simulated and measured S-parameters of the proposed back to back structure with different lengths (a) Short waveguide (SWG) and (b) Long waveguide (LWG)

4.4 Design of Novel DBAs Integrated with Proposed Low Loss Feeding Substrate Integrated Guide in the Sub-THz Band

In chapter 1, the advantages of wireless communications in the Sub-THz band (100-300 GHz) are mentioned and a need for the design of high gain and wideband antennas is presented. However, many designed antennas in the Sub-THz frequency band, e.g., on-chip antennas, suffer from low gain and radiation efficiency [84, 91, 92]. To improve their performance, dielectric resonator antennas can be employed in the design of Sub-THz devices as a possible choice due to high gain, wideband, and low loss advantages [32, 36, 92]. Recently, the integration of a DRA and photonic crystal waveguide operating in the frequency range of 311 to 331 GHz has been reported in [100], although the lack of a ground plane and supporting structure are drawbacks to the designed antenna. Developing an integrated antenna and low loss feeding mechanism can therefore be a promising solution for high frequency applications based on substrate integration technology [95].

First, an idea and the feeding mechanism of the antenna are introduced. A novel dielectric resonator antenna integrated with a low loss substrate integrated image guide in the Sub-THz band is then proposed. Finally, the design and manufacture of a dual-element, novel, high gain, dielectric based antenna integrated with the proposed low loss feeding mechanism at a frequency range of 240-270 GHz are presented.

To couple energy to a dielectric-based antenna, the feeding low loss waveguide operates at the fundamental mode of TE_{10} . By considering this fact and looking at the field distribution directions, the field distribution directions of the TE_{10} mode in the waveguide are like a HEM_{11} mode of the cylindrical dielectric resonator, as shown in Fig. 4.16. To take advantage of this, we create defects inside the guiding channel, and by doing so, the designed waveguide can excite a specific mode directly, and broadside radiation patterns can be expected. To show the design procedure, two feeding waveguides are considered, namely a substrate integrated image guide (SIIG) and the proposed substrate integrated guide, and the necessity to design a new waveguide proposed in the previous section is elaborated.

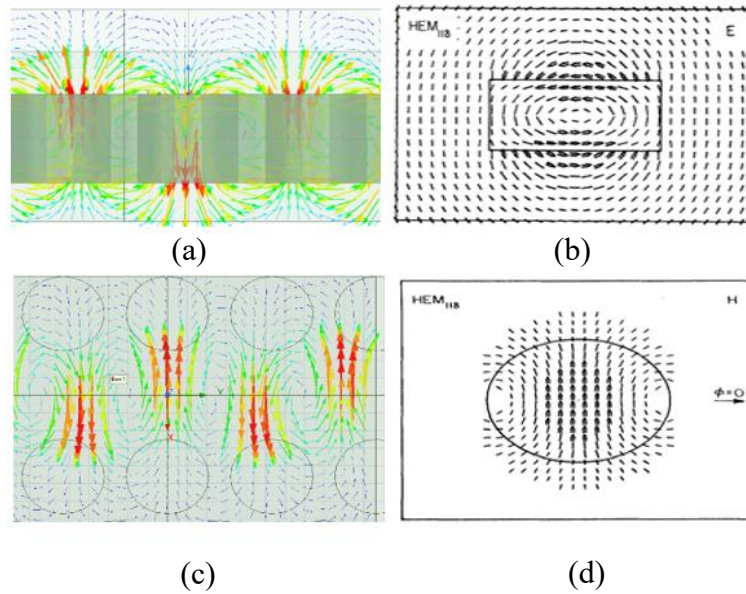


Figure 4.16 E field distribution of the (a) guiding channel and (b) HEM_{11} mode of a cylindrical dielectric resonator © [1984] IEEE [40]. H field distribution of the (c) guiding channel and (d) HEM_{11} mode of a cylindrical dielectric resonator © [1984] IEEE [40]

4.4.1 Design of a Novel Sub-THz Dielectric Resonator Antenna Fed by Substrate Integrated Image Guide (SIIG) Technology at 165 GHz

To show the importance and novelty of the design of the integrated DBA that will be presented in the next section, a Sub-THz dielectric based antenna fed by the substrate integrated image guide (SIIG) technology is presented and its disadvantages are described.

SIIG operates based on a total internal reflection mechanism. It includes a guiding channel of high dielectric constant in the middle, two symmetric air hole perforated regions on two sides of the guiding channel, and a supporting ground plane [87]. The proposed novel DRA is created by making a defect region along the guide channel of a SIIG structure. To couple the power from the rectangular waveguide WR5 to the SIIG guiding channel, a matching dielectric taper is employed. In addition, the use of tiny metal cylinders like pins at the end of the guiding channel results in the creation of an electric wall. Figure 4.17 shows the designed structure. As can be seen, an array of air holes are created in the HR-silicon film with a permittivity of $\epsilon_r=11.69$ and a resistivity of 10 k Ω -cm. To prevent the band gap effect that the Bragg condition may cause [87, 95], the dimensions of the periodic lattice constant and air hole diameter are taken into account. The physical

dimensions of the structure, namely the width (W), length (L), thickness (t), diameter of the air holes (d_1), central air hole diameters (d_2), central pin diameter (d_3), lateral pin diameters (d_4), periodic lattice constant (p), and taper length (L_t) can be found in Table 4.1.

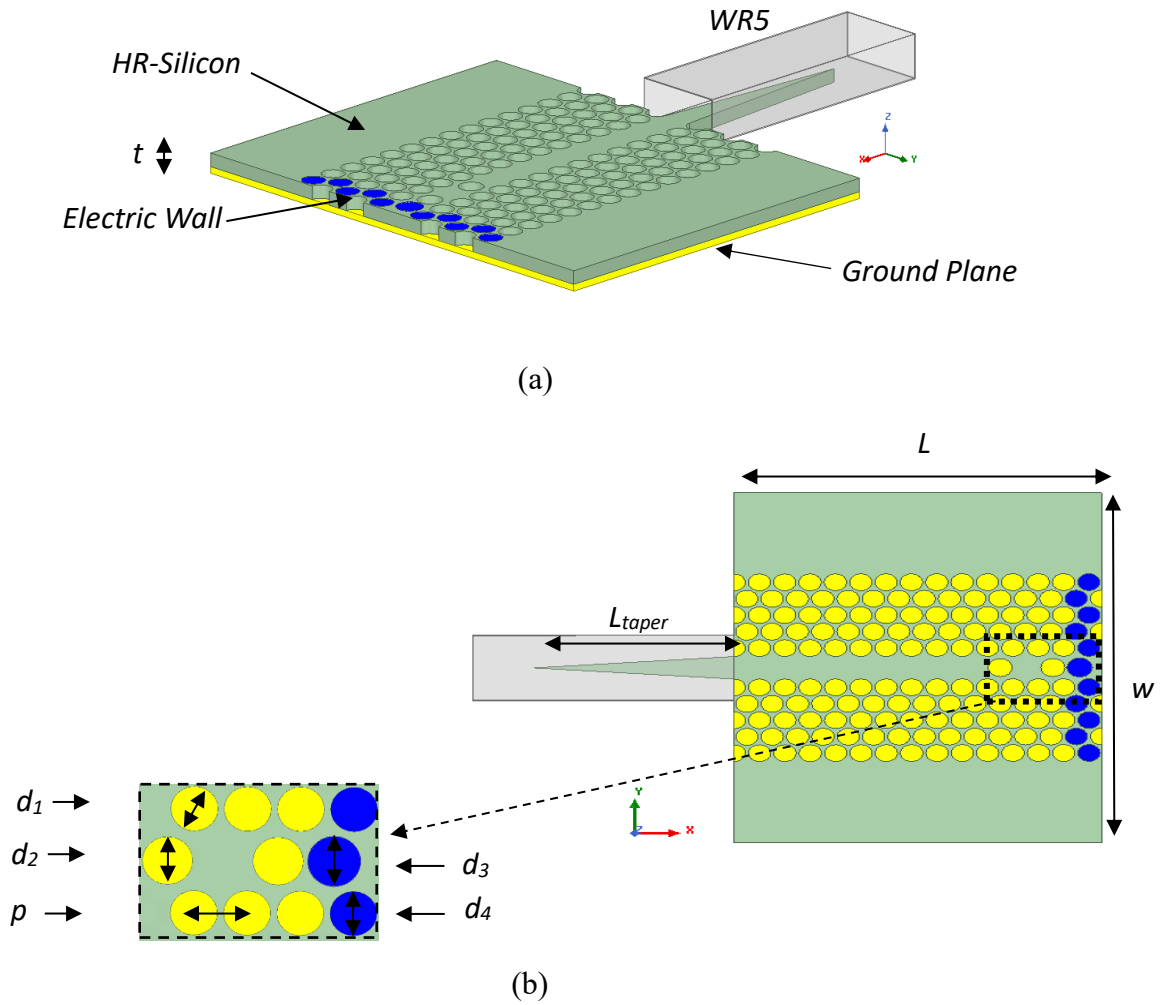


Figure 4.17 The geometry of the proposed integrated antenna (a) 3D perspective view and (b) Top view

Table 4.1 The parameter dimensions (in mm) of the designed DRA

Air holes	Pins	Substrate	Taper
$d_1=0.335$	$d_3=0.38$	$w=10$	$L_{taper}=4$
$d_2=0.36$	$d_4=0.335$	$L=5.5$	
$p=0.38$		$t=0.2$	

The performance of the designed antenna is investigated using High Frequency Structure Simulator (HFSS) software v2021. The S-parameter of the designed SIIG-DRA structure is shown in Fig. 4.18(a). The antenna resonance is at the frequency of 165 GHz and it covers a 21 GHz impedance bandwidth (12.6%). As Fig. 4.18 (b) shows, the magnetic field distribution inside the antenna confirms the excitation of the $HEM_{11\delta}$ mode at the resonance frequency of 165 GHz. Figure 4.19 (a) shows the antenna gain along the z-axis ($\theta = 0^\circ$). As can be seen, an average gain of the designed antenna is around 7.35 dBi within the wide range of frequencies (156-174 GHz), and the peak gain in the broadside direction is 8.5 dBi at the frequency of 163 GHz. To better clarify the reason why we need a new design configuration, the radiation pattern at the resonance frequency of 165 GHz in two different planes, the E-plane (XZ) and the H-plane (YZ), is illustrated in Fig. 19(b). The main beam of the H-plane radiation pattern is along the z-axis ($\theta=0^\circ$ direction). However, the E-plane radiation pattern is tilted and there are ripples on the radiation pattern that can be attributed to the asymmetric structure at the XZ-plane and the diffraction from the structure edges. Therefore, it is necessary to design a new feeding mechanism and configuration to overcome this problem.

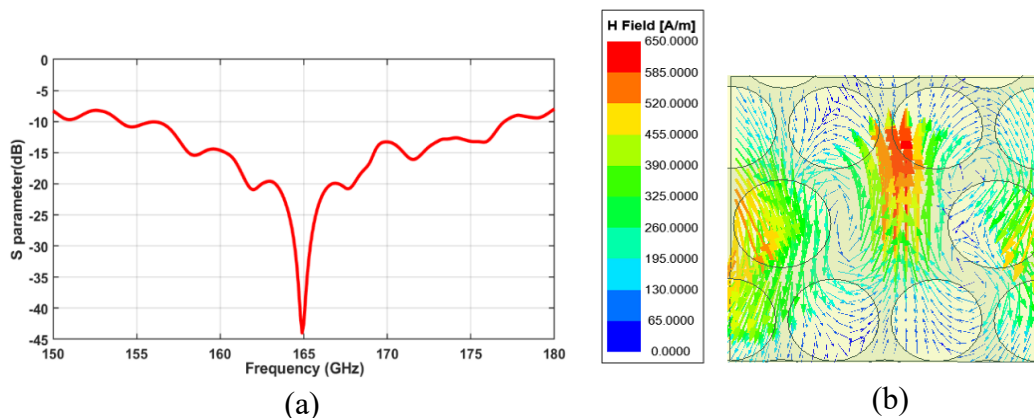


Figure 4.18 (a) Simulated S parameter of the proposed antenna and (b) H-Field distribution inside the DRA in the guiding channel of SIIG at the frequency of 165 GHz

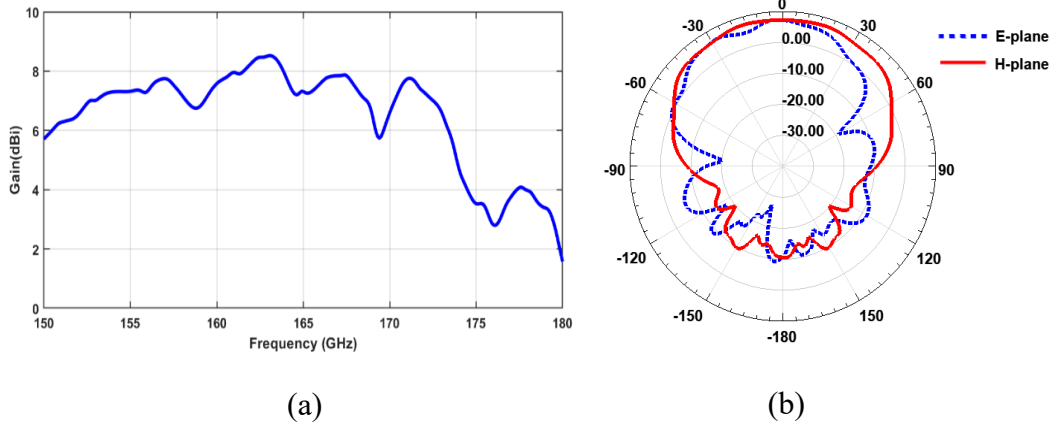


Figure 4.19 (a) Gain of the proposed antenna at broadside direction ($\theta = 0^\circ$), (b) Radiation pattern (gain) of the proposed antenna at resonance frequency of 165 GHz

4.4.2 Design of a Novel Dielectric Based Antenna Fed by the Proposed Feeding Technique in the Frequency Range of 230-270 GHz

In this section, to design a wideband and high gain antenna in the frequency range of 230-270 GHz, a dual element, novel, dielectric based antenna fed directly by the proposed substrate integrated waveguide with the same dimensions presented in the previous section is proposed. The designed structure is depicted in Fig. 4.20. It consists of two equal dielectric-based antenna elements created inside the guiding channels by making cavities, set close to each other, and excited by the E_z^{11} mode of the proposed dielectric waveguide. The proposed antenna design is completely symmetric. Therefore, symmetric radiation patterns are expected. Alumina with a permittivity of 9.8 and $\tan \delta = 0.0001$, and thickness of $t_a = 0.254$ mm is selected as a DBR structure. The alumina layer is supported by a RO4003C dielectric layer that has a permittivity of 3.55 and a thickness of $t_d = 0.305$ mm. The physical dimensions of the structure, namely width (W), and length (L), periodic lattice constant (p), and diameter of the air holes are 10 mm, 10.7 mm, 0.264 mm, and 0.2 mm, respectively. To prevent any internal coupling between the two ports, an aluminium wall is added between two dielectric based antenna (DBA) elements. The width ($W_m = 0.254$ mm) and length ($L_m = 7$ mm) of the aluminium wall are considered based on the fabrication possibility in Poly-Grames. In Fig. 4.20, two designations of 2x1y and 2x3y show a number of defects along the x-axis and y-axis. It should be noted that the defects along the guiding channels radiate, and radiating elements are shown in Fig. 4.20 as the main antenna.

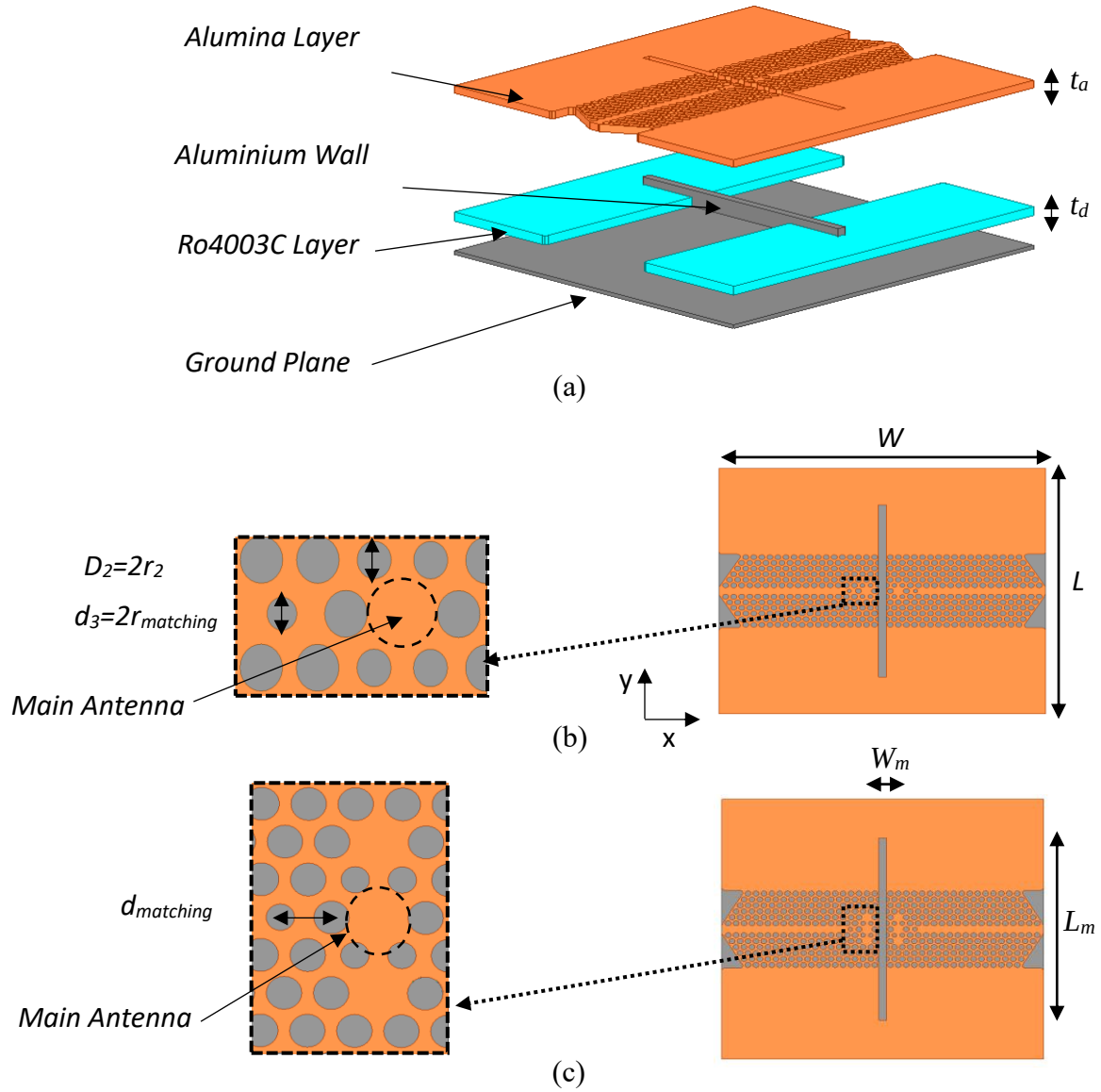


Figure 4.20 The geometry of the proposed dielectric based antenna (a) Diametric view, (b) 2x1y (top view), and (c) 2x3y (top view)

4.1.1.4 Simulation and Discussion

To show the performance of the proposed DBA, radiation characteristics namely impedance matching, mutual coupling, efficiency, gain, and radiation patterns of the antenna are investigated. To improve the impedance matching (-10 dB bandwidth), an air hole with the radius and distance of r_{matching} and d_{matching} is created. The dual element DBAs are fed by a differential feeding mechanism.

Here, the reflection coefficient of the DBA without using the matching air hole is investigated. After that, we demonstrate the effect of employing matching air holes on S-parameters. The dimension of r_2 is swept to study their effects on the antenna impedance matching. Figure 4.21 (a) shows the results. As shown, by decreasing the r_2 dimension, the reflection coefficient at lower frequencies is improved. It can be attributed to increasing the size of DBA. Here, the dimension size of $r_2=0.085$ mm is chosen based on the manufacturing limit. Then, the matching air holes are added on two sides of the metal wall with a radius of $r_{\text{matching}}=0.0635$ mm and a spacing of $d_{\text{matching}}=0.2143$ mm. As can be seen in Fig. 4.21 (b), the -10 dB impedance band width increased, as expected. To improve the impedance bandwidth, we remove four air holes on two sides of each guiding channel of a dual element DBA.

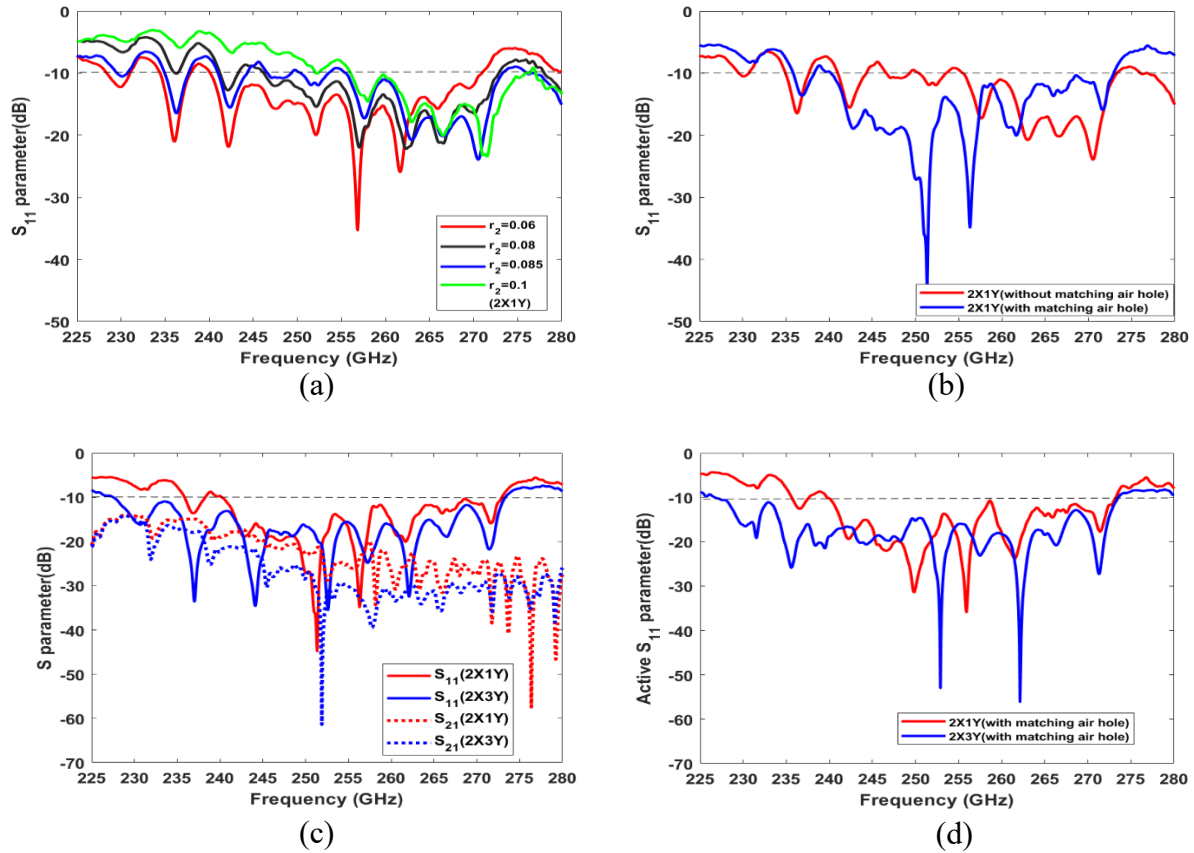


Figure 4.21 (a) S_{11} of the (2x1y) antenna for different dimensions of r_2 , (b) investigation of the air hole matching on S_{11} of the (2x1y) antenna, (c) S parameters of the designed DBA for two different configurations of (2x1y) and (2x3y), and (d) Active S parameters of the designed DBA for two different configurations of (2x1y) and (2x3y)

By comparing the obtained results, it is observed that the impedance bandwidth and the mutual coupling improved, as shown in Fig. 4.21 (c). The -10 dB impedance bandwidth was 18.3% at the frequency range of 227.5-273.3 GHz. Additionally, to examine the effect of ports on the reflection coefficients, the active S parameters are plotted in Fig. 4.21 (d). According to the results, the values of the active S parameter are acceptable.

Since the objective is the design of a high-performance DBA, the antenna efficiency is studied and illustrated in Fig. 4.22 (a). The average of the efficiencies is 91% and 93.6% for the two different configurations of (2x1y) and (2x3y), respectively. To examine the dual-element DBA performance, we studied the gain of the mentioned structures along the z-axis, i.e., the antenna boresight shown in Fig 4.22 (b). It was observed that the (2x3y) antenna configuration showed an improved gain in comparison with the (2x1y) antenna configuration. Hence, the gain increased almost all over the bandwidth. The average gain of the final design was 13.77 dB and 12.65 dB for the (2x3y) antenna configuration and the (2x1y) antenna configuration over the wide frequency range of 230-270 GHz, respectively.

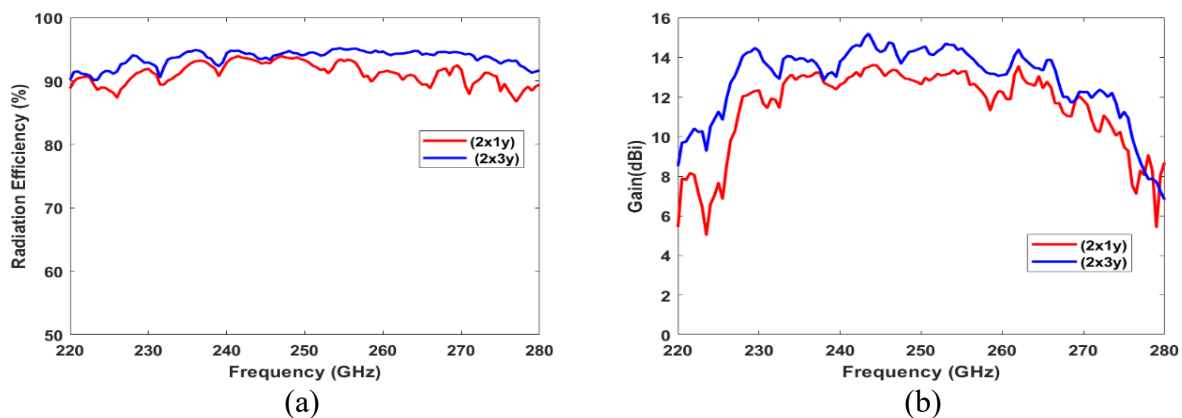


Figure 4.22 (a) Radiation efficiency and (b) gain (dBi) along the z-axis of two different configurations of (2x1y) and (2x3y) DBA

We studied the radiation patterns of the proposed DBA to gain a better understanding of the antenna performance. The sample radiation patterns are illustrated in Fig. 4.23 (a)-(d) in the xz-plane (E-plane) and the yz-plane (H-plane) for frequencies of 237 GHz, 244 GHz, 253 GHz, and 262.5 GHz. As can be seen, side lobe levels are less than -10 dB and the radiation patterns are appropriate and broadside. Since the designed structure is symmetric, the radiation patterns are symmetric as well.

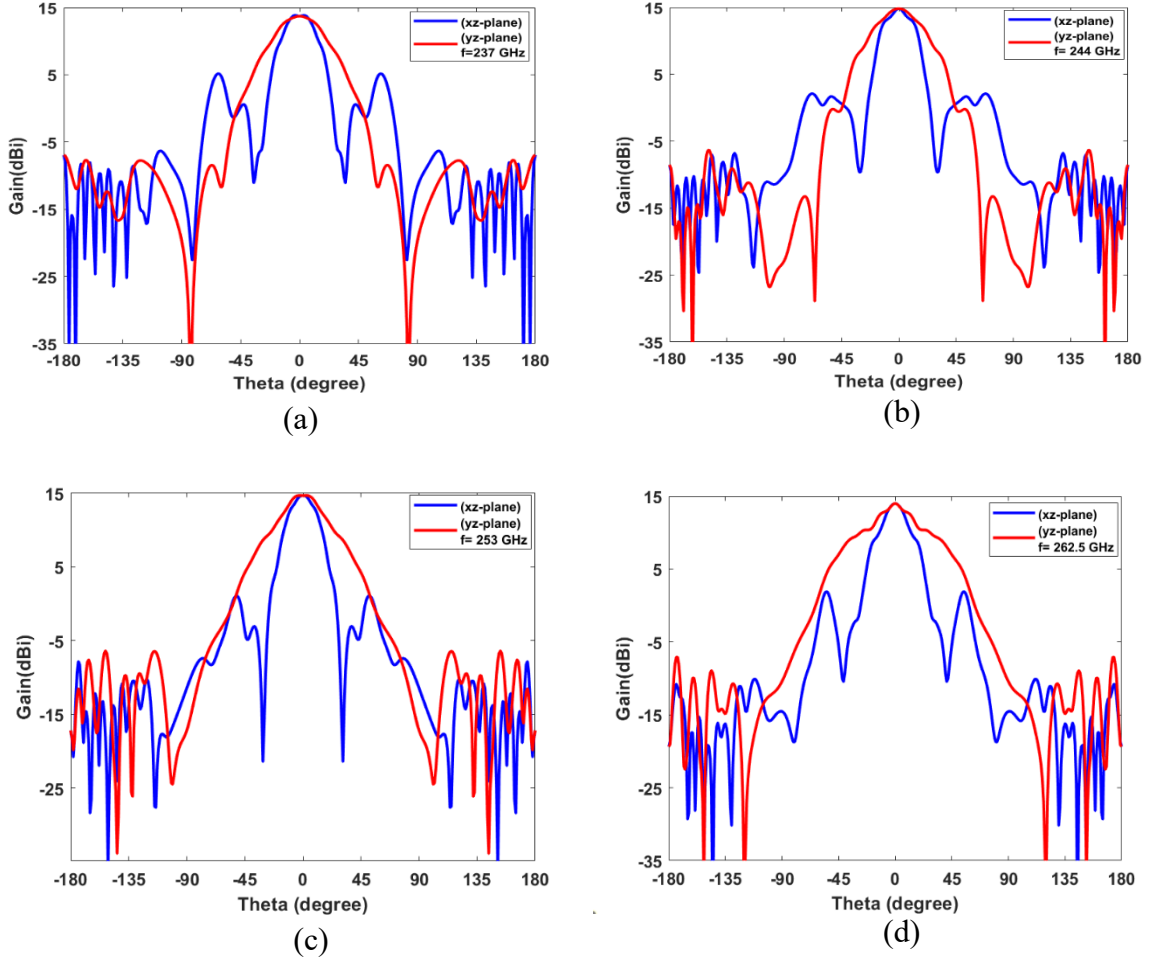


Figure 4.23 Simulated radiation patterns (gain) of the proposed dual-element DBA for the (2x3y) configuration, in the xz-plane (E-plane-blue line) and yz-plane (H-plane-red line) at the frequencies of (a) 237 GHz, (b) 244 GHz, (c) 253 GHz, and (d) 262.5 GHz

4.1.1.5 Fabrication

To measure S parameters and radiation patterns of the proposed DBA, a new aluminium fixure for the antennas was designed to connect the designed structure to extenders of the measurement setup. Since the designed structure includes two layers, an alumina waveguide and a dielectric supporting layer, metal screws were used to make them stable and avoid an air gap between layers. The geometry of the proposed DBA and photographs of the fabricated prototypes are shown in Fig. 4.24. The guiding channel of waveguides was excited by the rectangular waveguide WR3 with cross section dimensions $a = 0.8636$ mm by $b = 0.4318$ mm.

To assemble the antenna design, a larger slot was considered for the aluminium wall; therefore, an air gap was created between the central wall and alumina waveguide. To remove the mentioned air gap, an epoxy layer was applied in the air gap region. Furthermore, to reduce the effect of the fixture size, resulting from its large size as a scatterer, on the radiation patterns, WR3-band rectangular waveguides were extended.

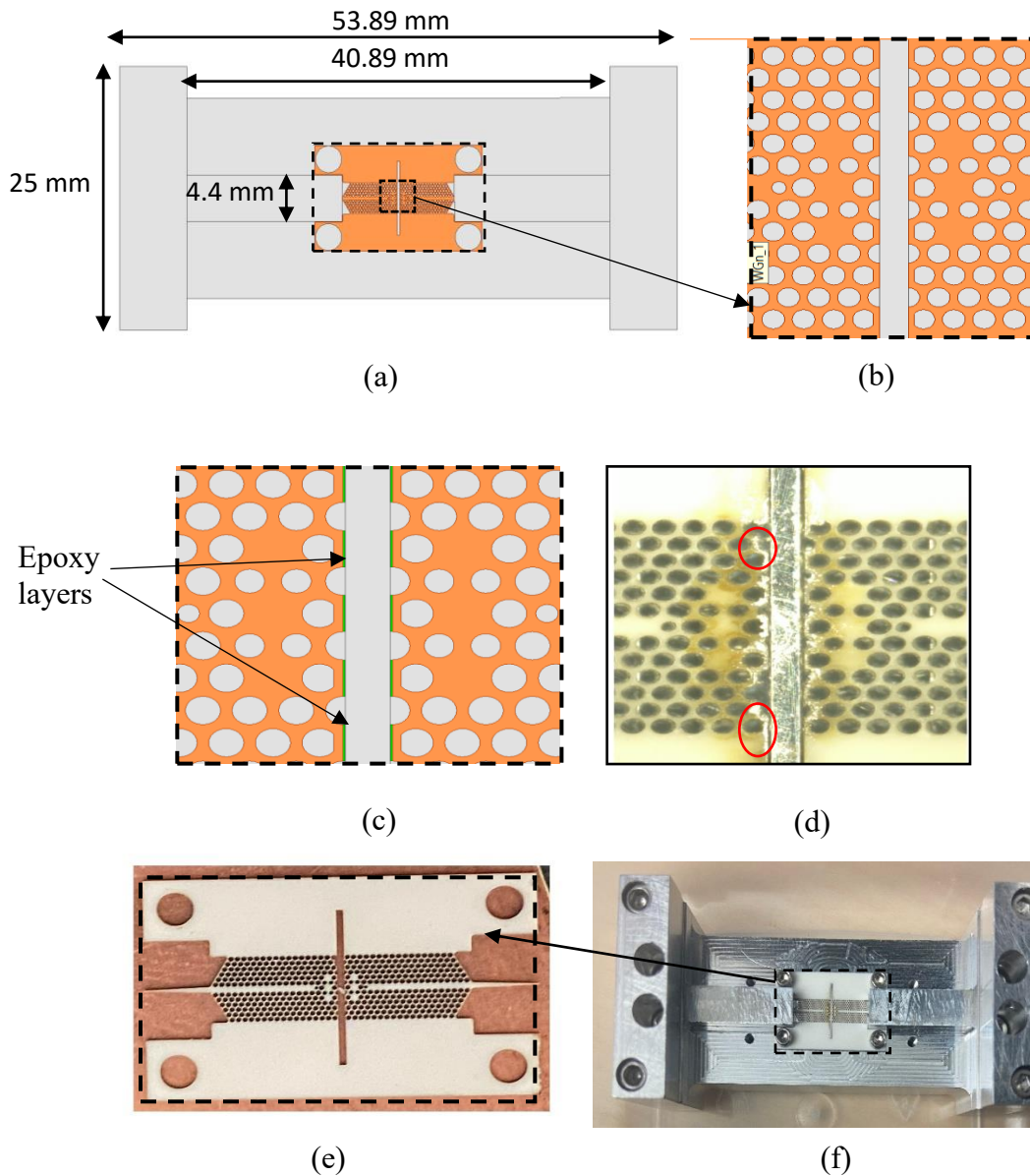


Figure 4.24 (a) and (b) The designed DBA for the fabrication, (c) and (d) The applied epoxy layers, (e) Photograph of the fabricated alumina layer, and (f) Photograph of the presented back-to- back DBA architecture

4.1.1.6 Fabrication Errors

A VNA equipped with VDI frequency extenders was used to measure S-parameters in the frequency range of 220-320 GHz, and the calibration was carried out. It was observed that the major part of the energy was reflected from port 1, and S_{11} and S_{22} parameters were not symmetric.

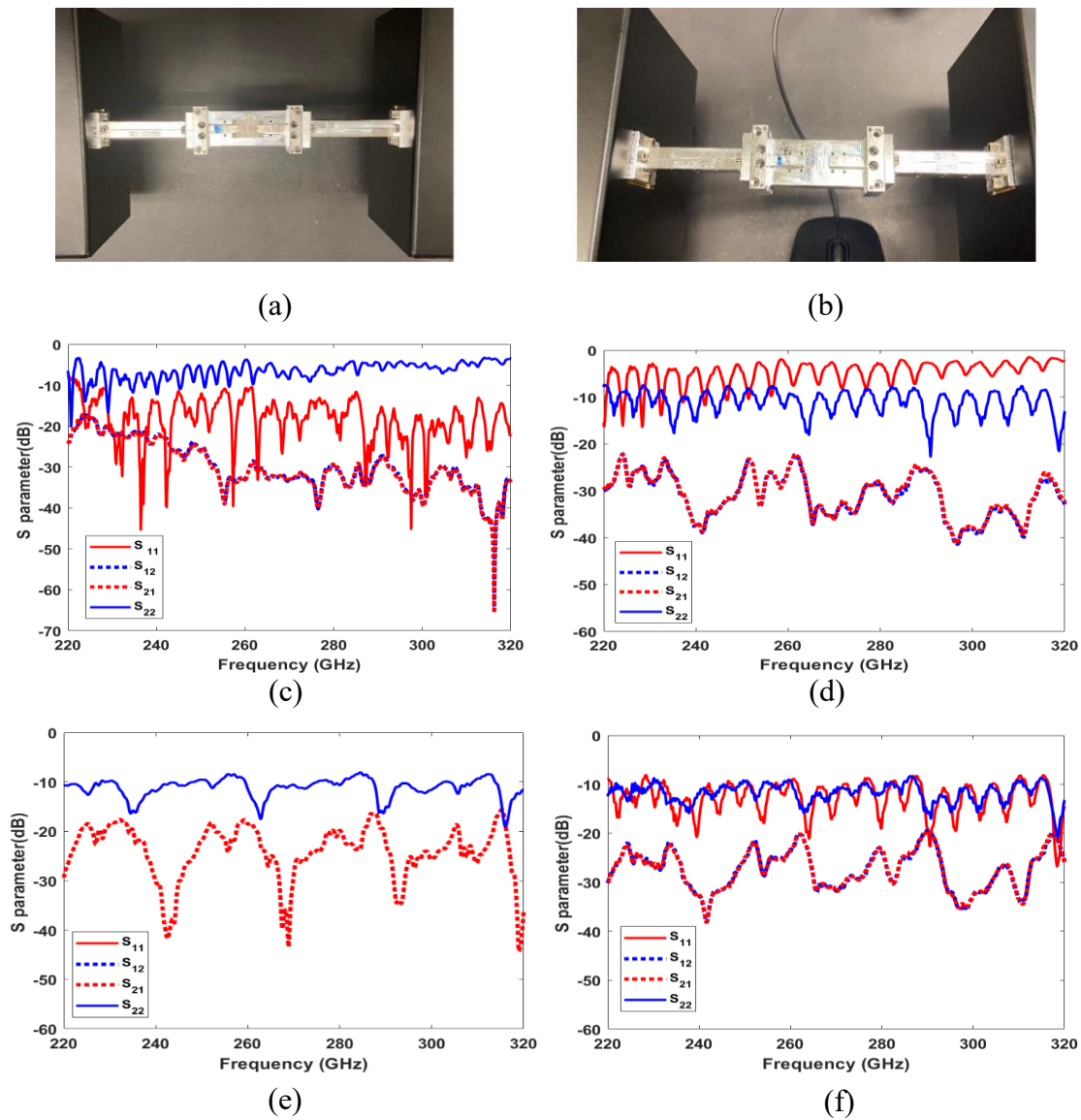


Figure 4.25 (a) and (b) Photographs of the aluminium fixture under testing with and without the DRBA, respectively. (c) and (d) Measured S-parameters for the aluminium fixture with and without the DRBA, respectively. (e) and (f) Simulated and measured S-parameters for the aluminium fixture without DBA, respectively

To find the error, we performed the S-parameters measurement after rotating the board (alumina layer) 180° as shown in Fig. 4.25 (a), but the issue persisted, as can be seen in Fig. 4.25 (c). The S parameter measurement of the fixture without the DBA shown in Fig. 4.25 (b) was then performed and the simulation was completed. The measured and simulated results are illustrated in Fig. 4.25 (d) and (e), respectively. As can be seen, the S_{11} and S_{22} parameters are not symmetric and the major part of the port 1 energy is still reflected. By comparing the measurement results, we can infer that the reflected energy results from the manufacturing issue of the fixture at port 1. Therefore, a new WR3 fixture was fabricated and the S-parameters were measured. The new measured results are shown in Fig. 4.25 (f). The comparison between the new measured S-parameters and the simulated results confirms that the WR3 fixture works and the issue is resolved. It should be mentioned that the epoxy layer between the metallic wall and alumina waveguide was not applied for these measurements.

4.1.1.7 Measurement Results and Discussion

In this section, first the S-parameters of the proposed dual-element dielectric-based antenna were measured. Next, the radiation patterns of the prototype were measured and the obtained results were compared with the simulation results. Figure 4.26 depicts the measurement system configuration. The designed antenna is connected to the measurement setup (VDI extenders) via the designed fixture. S-parameters were measured in the frequency range of 225-280 GHz using a Vector Network Analyzer (VNA). The proposed DBA under testing can be seen in Fig. 4.26 (b).

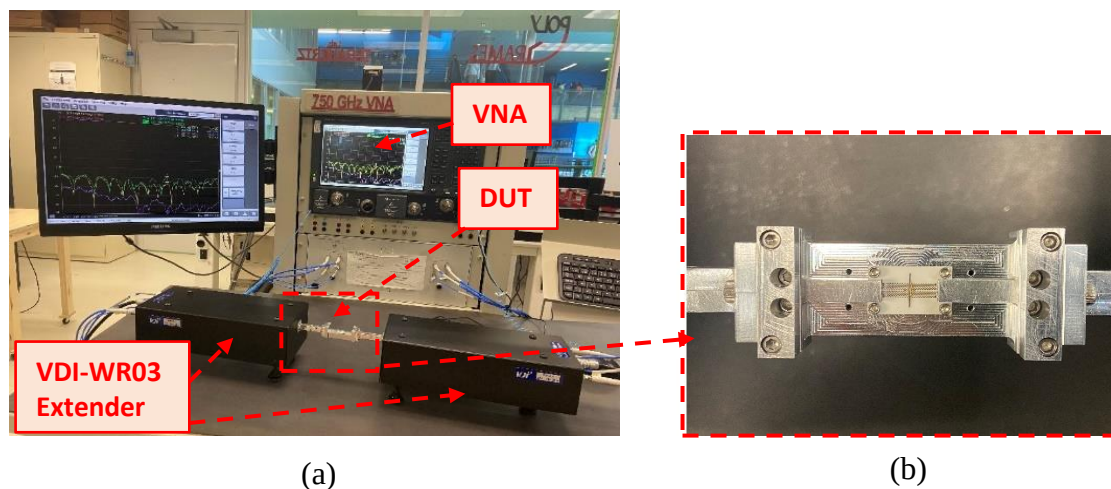


Figure 4.26 (a) THz measurement set-up (b) Top view of the proposed DBA under testing (DUT)

The S-parameters of the designed DBA were studied. Fig. 4.27 (a) shows the simulated and measured reflection coefficients. The reflection coefficient of port 2, S_{22} (blue line), is more than that of port 1, S_{11} (red line), which can be attributed to the fabrication error highlighted in Fig. 4.24 (d). There is also a little difference between the measured reflection coefficients of the designed fixture without the board (DBA), as shown in Fig. 4.25 (f). The measurement results show a proper -10 dB impedance matching and confirm the simulation results. The mutual coupling, S_{21} and S_{12} parameters, was excellent and the measured results confirm the obtained simulation results.

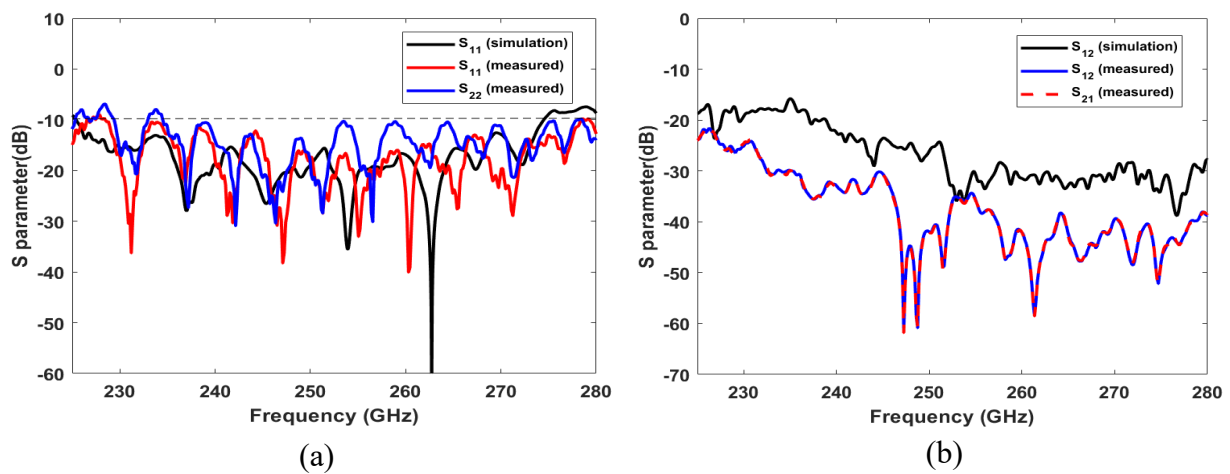


Figure 4.27 Simulated and measured (a) the reflection coefficient and (b) the mutual coupling for the proposed antenna

To obtain better insight into the performance of the designed DBA, the radiation patterns of the designed antenna for the obtained simulation and measurement results were studied and compared. It is worth noting that WR3-band rectangular waveguides were extended to overcome the inappropriate effect of the scattering radiation on the antenna performance. To analyze the radiation patterns of the designed DBA, the fabricated prototype was tested in the anechoic chamber at the Poly-Grames Research Center for far-field radiation patterns at three different frequencies. Figure 4.28 shows port 1 of the DBA prototype connected to the VDI and port 2 connected to the matching load. After doing the measurement, port 2 of the DBA prototype was connected to the VDI, and port 1 was connected to the matching load. Radiation patterns were measured in xz and yz planes with a resolution of 5 degrees. As illustrated in Fig. 4.28 (b), the chamber could only handle measurements from -30° to 210° and -90° to 90° in the xz-plane and yz-plane, respectively.

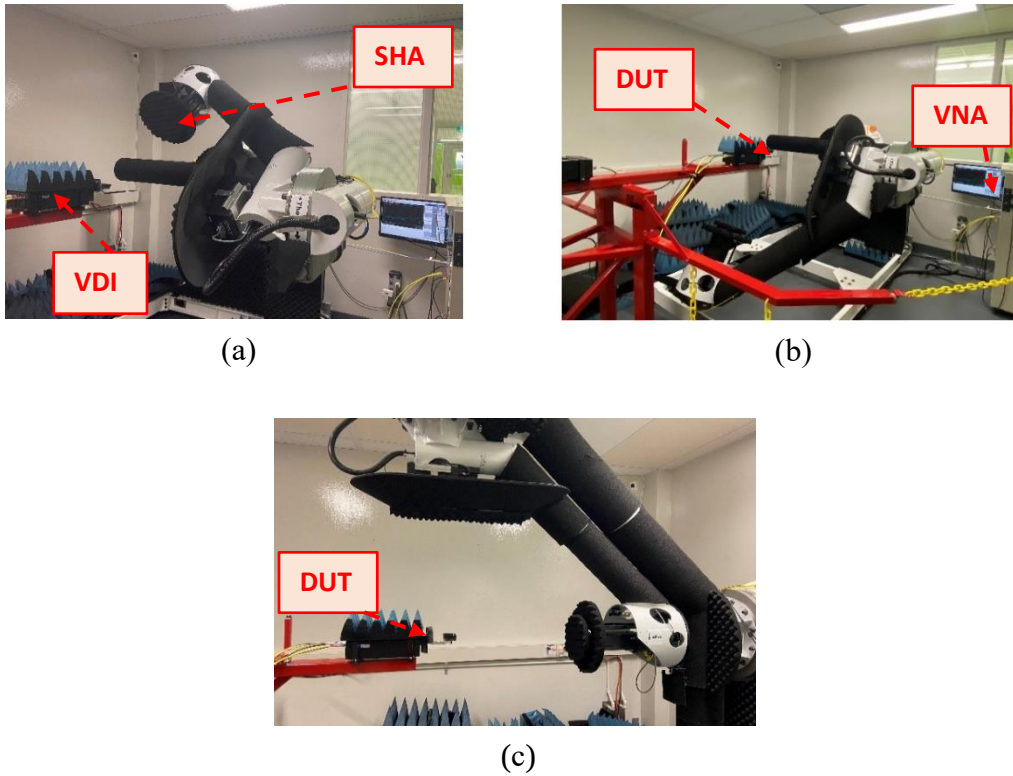


Figure 4.28 (a), (b), and (c) Photographs of the anechoic chamber and measurement set-up

The normalized radiation patterns of two ports of the designed DBA were measured and compared to the simulation results in xz -plane and yz -plane at frequencies of 247 GHz, 257 GHz, and 267 GHz. Since the structure is symmetric, the simulation radiation pattern of only port 1 of the DBA is considered for comparison purposes. As can be seen in Fig. 4.29, an excellent agreement was seen between the simulation and measured results. It should be noted that it was not possible to measure the DBA radiation patterns from $\theta = -150^\circ$ to $\theta = -30^\circ$ due to the red framework obstacle of the anechoic chamber in the xz -plane, as can be seen in Fig. 4.29 (a), (c), and (e). Additionally, to mitigate the effect of antenna set-up misalignment, an alignment adjustment factor is considered. There are a few ripples on the measured radiation pattern of the DBA port 2 that could be attributed to the fabrication tolerance, including the assembly of the alumina layer, the spacing between wall and alumina layer, the antenna set-up misalignment, and the merging of some air holes in the designed waveguide excited by port 2, as shown in Fig. 4.24 (d).

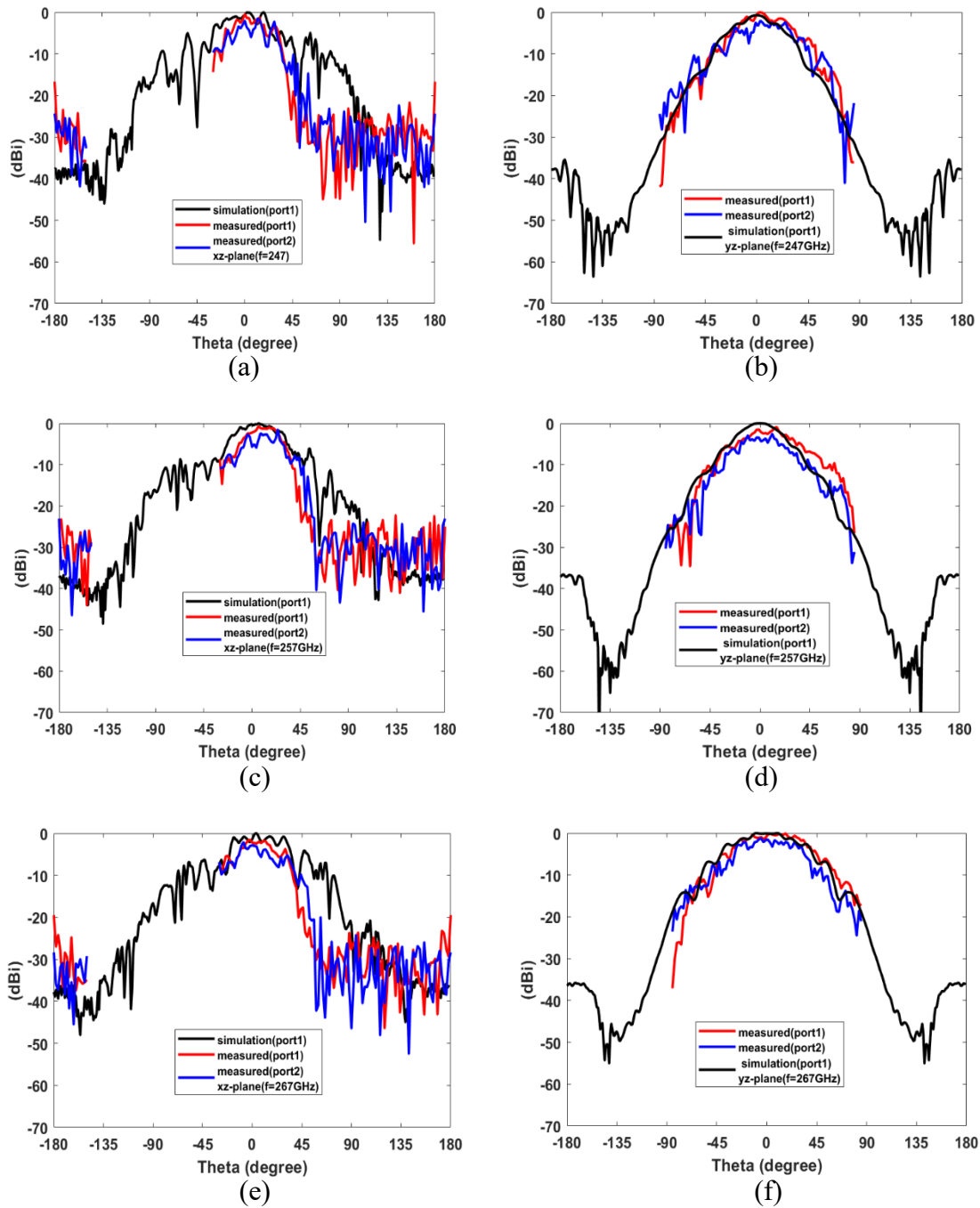


Figure 4.29 The simulation and measured radiation patterns of the designed DBA for three different frequencies of (a) and (b) 247 GHz, (c) and (d) 257 GHz, (e) and (f) 267 GHz in the xz-plane and yz-plane

The radiation pattern combination obtained from the simulation results is illustrated in Fig. 4.30. The proper side lobe level and high gain in the z-axis direction were achieved. The simulation antenna gains are 14.24 dBi, 14.4 dBi, and 12.21 dBi at the frequencies of 247 GHz, 257 GHz, and

267 GHz, respectively. In addition, there is low cross polarization within the main beam. It should be mentioned that the simulation cross polarization radiation patterns in the xz-plane are not shown since the values are less than -60 dBi.

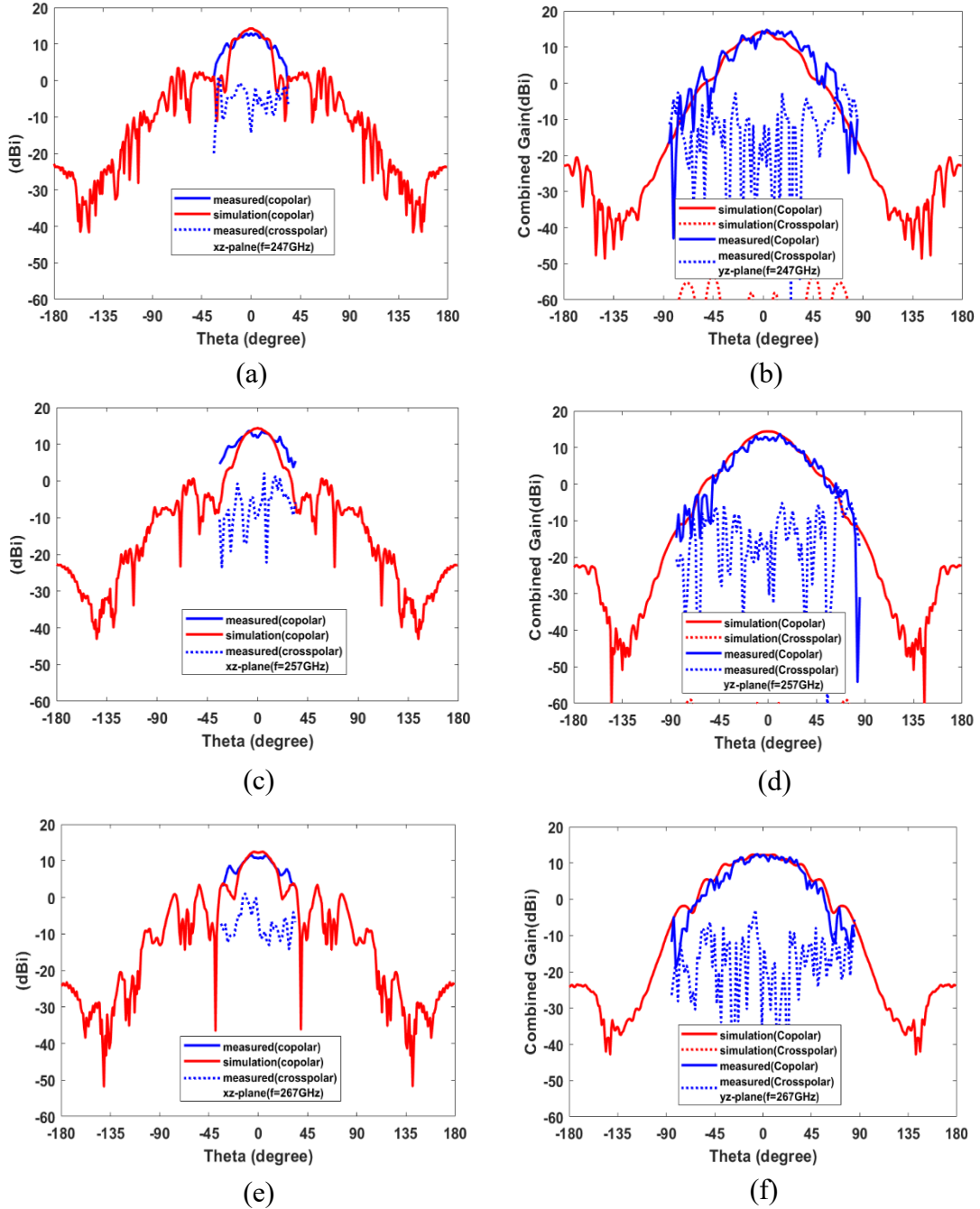


Figure 4.30 Simulated and measured radiation patterns (combined gain) of the proposed dual-element DBA for the (2x3y) configuration at three various frequencies of (a) and (b) 247 GHz, (c) and (d) 257 GHz, (e) and (f) 267 GHz in the xz-plane and yz-plane

A table of comparison with the most recent dielectric-based antennas operating in the Sub-THz band is presented in Table 4.2. It is obvious that the proposed DBA can provide good performance in the Sub-THz frequency band namely high gain and proper impedance bandwidth.

Table 4.2 A comparison summary between previous works and the proposed DBA

Ref.	Frequency Bands (GHz)	BW(GHz)	Realized Gain /Gain	Antenna Type
[105]	186-205.5 GHz	9.96%		Nano DRA
[35]	172 to 185 GHz	7.3%	7-8 (Gain)	Spherical DRA
[94]	110-130 GHz	-	16 (Gain)	Tapered antenna
[100]	311 to 331 GHz	6%	10.6 (Gain)	DRA fed by photonic crystal waveguide
[102]	1.483 to 1.535 THz	3.46%	7.4 (Gain)	DRA fed by photonic crystal waveguide
[106]	130 GHz	10%	7 (Gain)	On-chip DRA (4×1 array)
[107]	319 to 360 GHz	12%	10 (Realized Gain)	On-chip antenna (Yagi like)
[108]	331-356 GHz	7.3%	7.9(Realized Gain)	On-chip DRA
This work	233.5-278.3 GHz	17.5%	13 (Gain)	DBA-fed by the proposed waveguide

4.5 Conclusion

In this chapter, a newly developed substrate integrated waveguide based on SOG technology was proposed, operating in the Sub-THz band (140-180 GHz) range of frequencies. To design a feeding mechanism for dual-element dielectric-based antennas operating in the frequency range of 230-270 GHz, a newly developed substrate integrated alumina guide was proposed. The designed structure showed proper simulation results (insertion loss of less than -1). The presented waveguide was fabricated based on laser micromachining technology to create the perforation air holes. Additionally, a new metal fixture was designed and fabricated using CNC machining technology

to connect the designed waveguide to extenders of the measurement setup. The measurement results were acceptable (the insertion loss was between -2 dB and -4 dB).

A novel dual-element dielectric-based antenna fed by the proposed substrate integrated alumina guide was presented and studied. The design procedures along with the simulation results were discussed in detail. The simulation results indicated the high radiation efficiency (average of 93.6%) and gain (average of 13.77 dBi) with the broadside radiation patterns in the Sub-THz band over the frequency range of 230-270 GHz. Subsequently, the dual-element, dielectric-based antenna was manufactured and successfully tested. The measurement results indicated the -10 dB impedance bandwidth of 17.5 %. The proposed DBA could be a good candidate for short range wireless communications in the Sub-THz band.

CHAPTER 5 CONCLUSION AND FUTURE WORKS

Based on the results and discussions from the previous chapters, we reached the conclusion in this chapter, and future works are mentioned.

5.1 Conclusion

In this thesis, we focused on the design of high gain and wideband dielectric antennas in the millimeter wave and Sub-THz bands. In considering the huge unallocated spectrums in these frequency bands, the growing demand for high data rates, and the advancement of fabrication technology, the antenna design operating at the Sub-THz frequency could be a potential solution. A perforated cylindrical dielectric resonator antenna was therefore proposed to cover the unlicensed frequency range of 57-64 GHz. A 6 dB power divider with the same phases at output ports was then presented based on SIW technology to excite the cylindrical DRAs array.

Two different substrate-integrated waveguides were presented at the frequency ranges of 140-180 GHz and 230-300 GHz. The first one was designed based on silicon-on-glass (SOG) technology. The second one, a developed dielectric integrated waveguide, was proposed and fabricated as a feeding mechanism of the dielectric-based antenna using the laser micromachining to create the perforation air holes. The designed structure operated on the frequency range of 220-280 GHz with the appropriate insertion loss between -2 dB and -4 dB.

Finally, a novel, dual-element, dielectric-based antenna fed by the proposed substrate integrated alumina guide was presented and studied. The design procedures along with the simulation results were discussed in detail. The simulation results indicated high radiation efficiency (average of 93.6%) and gain (average of 13.77 dBi) with the broadside radiation patterns in Sub-THz band over the frequency range of 230-270 GHz. Moreover, it was fabricated and successfully tested. The major contribution of this thesis is the design of a high gain, small size, wideband, completely embedded antenna in a waveguide system in the Sub-THz band. Therefore, the proposed DBA could be a good candidate for short range wireless communications in the Sub-THz band.

5.2 Future Works

The scope of future work is vast. The goal of this research project was the design of a high gain antenna in the Sub-THz band. Future works that could be attractive for future wireless communication in the Sub-THz and THz bands are presented.

1-Design the large array using the proposed dielectric based antenna fed by dielectric waveguide-power divider as a feeding mechanism for achieving a higher gain antenna that can be proper for short-range THz wireless communication, and the study of the circularly polarized dielectric-based antenna in the Sub-THz frequency range.

2-Design and modal analysis of the high gain DRA array in a 4×4 element configuration at various frequency ranges of 141-148.5 GHz, 151.5-164 GHz and 209-226 GHz for wireless communications and comparing the simulation and measurement results.

3- Design the series-feed array configuration of the presented perforated dielectric resonator antenna to resonate at different frequencies with the purpose of bandwidth improvement..

4-Design the power divider with different phases at the output ports as a feeding mechanism for the design of the dielectric-based antenna with the purpose of beam steering.

5-Modal analysis of the proposed dielectric waveguide and developing a theoretical formula for achieving the propagation and attenuation constants. This study has been started, and will continue until a theoretical formula is developed.

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APPENDIX A LIST OF PUBLICATIONS

- 1-E. Rahmati and M. S. Sharawi, "Design of a Novel Sub-THz Dielectric Resonator Antenna Fed by Substrate Integrated Image Guide (SIIG) Technology" *2023 IEEE International Symposium on Antennas and Propagation and USNC-URSI Radio Science Meeting (AP-S/URSI)*, Accepted.
- 2-E. Rahmati, D. N. Aloï, and M. S. Sharawi, "60 GHz Perforated Dielectric Resonator Antenna Fed by Substrate Integrated Waveguide" *2023 IEEE International Symposium on Antennas and Propagation and USNC-URSI Radio Science Meeting (AP-S/URSI)*, Accepted.
- 3-E. Rahmati and M. S. Sharawi, "Substrate Integrated Dielectric waveguide for Sub-THz and Terahertz Applications", in preparation.
- 4-E. Rahmati and M. S. Sharawi, "Design of a Novel Dual-Elements Dielectric Based Antennas for Sub-THz applications," in preparation.