

Article

Three-Dimensional Printed, Dual-Band, Dual-Circularly Polarized Antenna Array Using Gap Waveguide Technology

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Abstract: A gap waveguide (GW)-based 4×4 dual-band, dual-circularly polarized antenna array is proposed. The antenna is composed of two stacked layers which are responsible for right-handed circularly polarized (RHCP) radiation in K-band and left-handed circularly polarized (LHCP) radiation in Ka-band. Each stacked layer consists of a GW power distribution network to excite 16 radiating units from a single input. A waffle grid is mounted on top of the array structure to increase effective aperture areas and reduce grating lobes. A prototype was fabricated using stereo lithography appearance (SLA) 3D printing and metallization technology, which significantly decreased the fabrication cost and complexity. Measurement results for the antenna prototype demonstrate the operating bandwidths of 19.91–20.52 GHz in K-band and 28.26–29.19 GHz in Ka-band, over which the reflection coefficients of < -10 dB and axial ratios of < 3 dB are achieved. The prototype provides peak gains of 20.1 and 20.3 dB with total efficiencies of $> 90\%$ in the two respective bands. The combination of dual-band, dual-circularly polarized capability, high gain, high efficiency, lightweight, low cost and compact size renders the proposed design a potential candidate for commercial millimeter wave communication applications.



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Keywords: 3D printing technology; antenna array; dual-band dual-circularly polarized; gap waveguide (GW)

1. Introduction

Circularly polarized (CP) antennas have advantages in resisting rain attenuation, suppressing multipath interferences, and receiving arbitrary linear polarization direction. Therefore, CP antennas have been widely used in the field of wireless communication and global positioning [1–3]. Furthermore, dual-band antennas minimize the requirement of using two antennas for two bands. The dual-band dual-CP antenna, which keeps polarization of the two bands opposite to each other, further enhances the isolation between two ports [4,5]. The antennas, realized by using printed circuit board (PCB) technology, is an attractive alternative owing to the benefits of low profile, inexpensive cost, and high conformality [6]. In [7], dual-band, dual-CP antenna arrays were implemented with single-layer PCBs. Ref. [8] presented a 2×2 magneto-electric dipole antenna array that realizes dual-band, dual-CP operation over a wide bandwidth. In [9], a 4×4 dual-band, dual-CP antenna array using an artificial magnetic conductor (AMC) ground and having a unidirectional radiation pattern was proposed. However, dielectric losses of those PCB-based antennas are non-negligible in the K- and Ka-bands, resulting in higher insertion losses and lower radiation efficiencies, especially for large-scale antenna arrays.

Compared with PCB-based antennas, full-metal antennas have advantages of low loss, high efficiency, and large power capacity. Slotted waveguide antennas have been demonstrated with high gains and efficiencies in the millimeter-wave band [10–12]. However, the involvement of high-precision computer numerical control (CNC) fabrication process significantly increases the cost. Additionally, potential fabrication error leading to

the formation of air gaps can result in field leakage, which decreases the efficiency of the antenna. To reduce the sensitivity to fabrication error and to simplify the CNC machining process, gap waveguide (GW) technology has been introduced and developed. The operating principle, fabrication, and applications of GW components have been explored in various studies [13–17]. GW-based antennas can be fabricated into different layers and then assembled without the need for high precision welding, which significantly reduces the implementation cost [18]. Various GW-based antennas with full metal construction have been reported to provide high efficiencies and high gains [19–23]. In [19], a GW-based, 4×4 Ka-band CP array is presented with a maximum efficiency of 98.0% and a maximum gain of 22 dB. In [20], an 8×8 Ka-band dual-polarized array based on GW technology was proposed. The measured gain and efficiency of the antenna are over 27.5 dB and 75%, respectively. This dual-polarized design was then combined with a hybrid coupler feeding network to realize dual-CP operation [21]. A few published works can be found on GW-based high-efficiency antennas with dual-band characteristics. In [22], a K-/Ka-band linearly polarized (LP) antenna array was proposed. The antenna provides a 25.5 dB gain with efficiency over 84% in K-band, and a 29.5 dB gain with efficiency over 88% in Ka-band. So far, dual-band, dual-CP antennas with metallic GW structures have rarely been reported. It is worth mentioning that a very preliminary approach to a 2×2 dual-band, dual-CP array antenna was proposed and numerically studied in [22]. However, the limited aperture size and small unit number results in a low directivity. The current work aims to develop a much more evolved version of that concept by addressing technical challenges of increasing the aperture size of the GW-based dual-CP array, finding a suitable fabrication process, and comprehensively investigating the antenna array's performance through both simulations and experiments.

In this paper, a GW-based, 4×4 dual-band, dual-CP antenna array is presented. The shared aperture cylinder cavity antenna array makes it possible to realize dual-CP radiation in a limited antenna aperture. The units of the proposed antenna array are excited by a feeding network consisting of both groove gap waveguides (GGWs) and ridge gap waveguides (RGWs). The modes and impedance transition of the feeding networks are carefully designed to guarantee the stability of the phase and amplitude distribution for each antenna unit. The antenna was built using stereo lithography appearance (SLA) 3D printing and metallization technology, which greatly decreases the difficulty and cost of the fabrication process [24,25]. The number of array elements can be periodically expanded to conveniently construct a larger-scale antenna array.

The paper is organized as follows. In Section 2, the design and optimization processes of the GWs; the dual band, dual-CP antenna unit; and the feeding network are discussed. Section 3 presents and discusses the simulated and measured results. Finally, the conclusion is drawn in Section 4.

2. Antenna Design Procedure and Methods

Figure 1 illustrates the three-dimensional structure of the proposed 4×4 dual-band, dual-CP antenna array. The Ka-band left-handed circularly polarized (LHCP) array is at the bottom layer of the antenna structure and the K-band right-handed circularly polarized (RHCP) array is stacked above the Ka-band array. A waffle grid is placed on top of the whole array structure to increase the effective area of apertures and reduce grating lobes. Two waveguide transitions were designed to connect the input ports of the K- and Ka-band arrays to corresponding standard waveguide interfaces to facilitate antenna measurement and system integration.

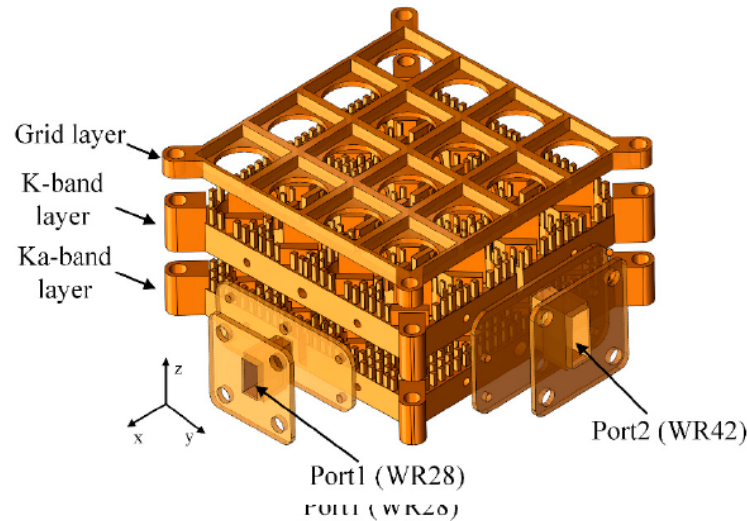


Figure 1. Illustration of the 3D structure of the proposed 4×4 dual-band, dual-CP antenna array.

2.1. Operating Principle and Simulation of the GWs

A GW is typically implemented by placing a metallic plate parallel to a high-impedance textured surface acting as an artificial magnetic conductor (AMC). When the spacing between the two parallel plates is smaller than a quarter of a wavelength, an electromagnetic band gap (EBG) is formed that forbids the propagation of all parallel-plate modes inside the region between the two plates. The textured surface usually consists of periodic metal pins to create a high-impedance boundary to realize the AMC condition over the operating frequency band [26]. If a certain area of the textured surface is replaced by a ridge or a groove, which is then surrounded by the metallic pins, wave propagation is restored and confined in the region above that modified area. This effectively creates a track for guiding the wave with minimum lateral leakage while providing the unique advantage of a contactless structure.

To design the GW components with the desired operating frequency ranges, we used the eigenmode solver of CST Studio, commercial full-wave simulation software, to obtain the dispersion diagrams for unit cells of the EBG structures. It can be seen from Figure 2 that the stopbands emerged in the Ka-band and K-band for the two respective parallel-plate structures. Optimized dimensions of these GW units were found to be $pa_1 = pa_2 = 0.9$ mm, $ph_1 = 2.4$ mm, $ph_2 = 3.1$ mm, $ag_1 = ag_2 = 0.1$ mm, and $d = 1.9$ mm.

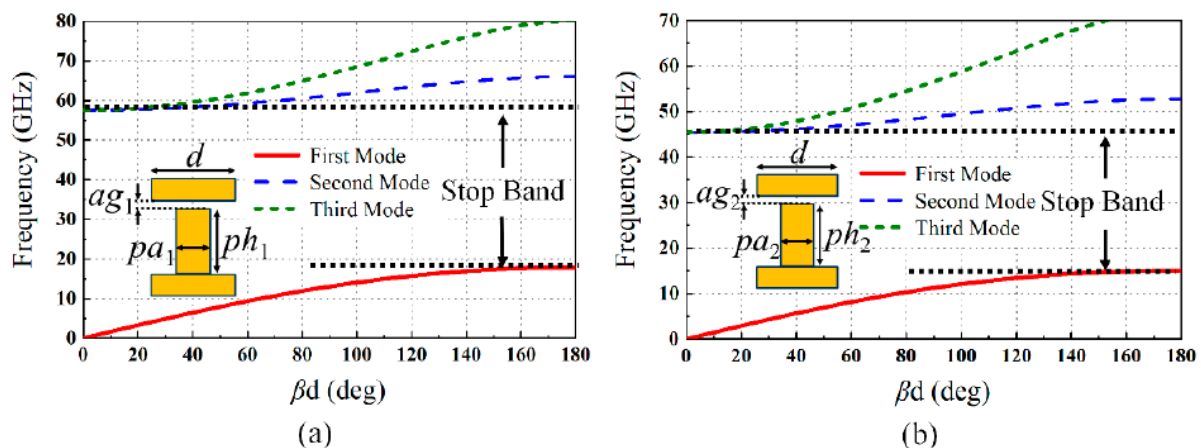


Figure 2. Simulated dispersion diagrams of the two GW unit cells. (a) Ka-band layer. (b) K-band layer.

2.2. Design of the Dual-Band, Dual-CP Antenna Unit

Figure 3 shows the 3D structural view of a CP antenna unit together with cross-sectional views of its Ka- and K-band cavities. The K-band cavity is stacked above the Ka-band cavity to share the common antenna aperture. The input ports of the antenna unit are on the sides of the two cavities. Each cavity has a cylindrical shape with two chamfered edges. The chamfered edges are created by placing two parallel metallic blocks that make an angle of θ with the feeding channel (see Figure 3). Owing to the perturbation effect of these angled blocks, two orthogonal modes of the cavity are excited with the same amplitude and a 90° phase difference. A 2.6 mm high cylindrical post with a radius of r_1 is placed at the center of the lower cavity to optimize the impedance of the Ka-band units.

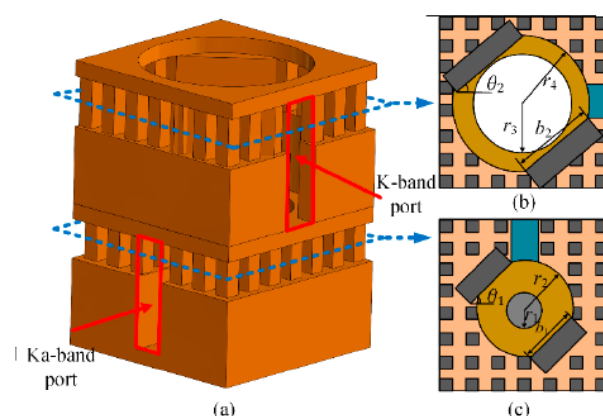


Figure 3. Illustration of an antenna unit. (a) 3D view of the antenna unit showing the stacked structure. (b) Sectional view of the K-band cavity. (c) Sectional view of the Ka-band cavity. The parameters of the antenna unit follow: $r_1 = 1.4$ mm, $r_2 = 3.3$ mm, $r_3 = 4.75$ mm, $r_4 = 5.4$ mm, $b_1 = 3.44$ mm, $b_2 = 6.4$ mm, $\theta_1 = 46^\circ$, and $\theta_2 = 37^\circ$.

The influence between the K- and Ka-band antennas is discussed as follows. For the K-band antenna, the Ka-band cavity can be regarded as a total reflection boundary. Hence, the radiation performance of the K-band antenna is not much affected by the Ka-band cavity below it. For the Ka-band antenna, the K-band cavity operates as a circular waveguide transmission line. The angled blocks of the K-band have some effect on the phases of the two orthogonal electric field components. However, this can be compensated by optimizing the dimensions and positions of the two angled blocks.

The field distributions inside the Ka- and K-band cavities at different time steps within one period T are shown in Figures 4 and 5, respectively. It can be seen that the E-field vectors are periodically rotated by 90° after each $T/4$ interval in the respective frequency band inside both cavities. The electric field vectors rotate clockwise at Ka-band and counterclockwise at K-band, which means LHCP and RHCP are realized at different frequency bands.

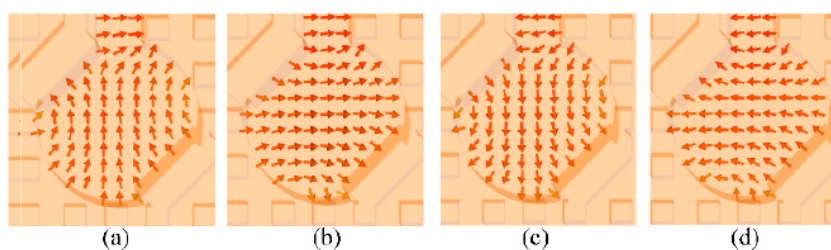


Figure 4. Electric field vectors inside the Ka-band cavity at different time steps within one period T . (a) $t = 0$. (b) $t = T/4$. (c) $t = T/2$. (d) $t = 3T/4$. The direction of wave propagation is coming out of the page.

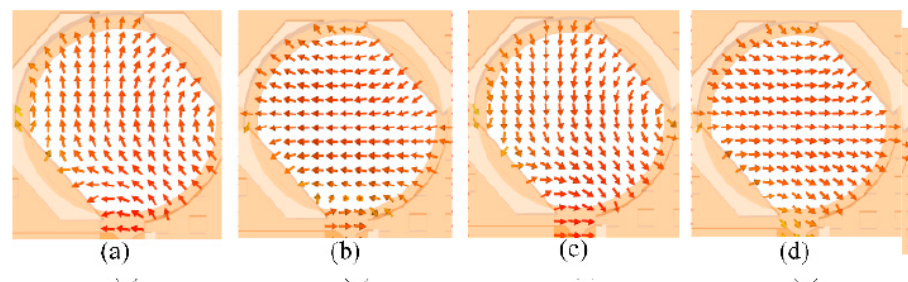


Figure 5. Electric field vectors inside the K-band cavity at different time steps within one period T . (a) $t = 0$. (b) $t = T/4$. (c) $t = T/2$. (d) $t = 3T/4$. The direction of wave propagation is coming out of the page.

2.3. Design of the 1 to 16 Feeding Networks

E-plane GGW corporate topologies provide compact feed networks, which were already used for a 4×4 slot array in [27]. However, E-plane GGW in 1-to-2 power distribution structure can cause output phase inversion, such as the input–output relationship of a magic tee in the E plane. Hence, an additional 180° phase shift is required to compensate for the phase difference and realize the broadside beam, which makes it difficult to further scale-up the array. On the other hand, implementing the feed network with RGWs can avoid such phase inversion [28]. However, RGW-based power dividers require additional impedance matching structures that occupy more space. These problems can be conveniently addressed when GGWs and RGWs are combined as part of the same distribution network [29]. Such an approach is optimal to feed arrays with symmetric and compact topologies and therefore was adopted to design the feeding network for the proposed antenna array.

Figure 6 shows the respective top views of the K- and Ka-band layers of the antenna array along with their feeding networks. We highlighted GGWs in blue and RGWs in red. Sixteen antenna units are fed by the paralleled feeding network.

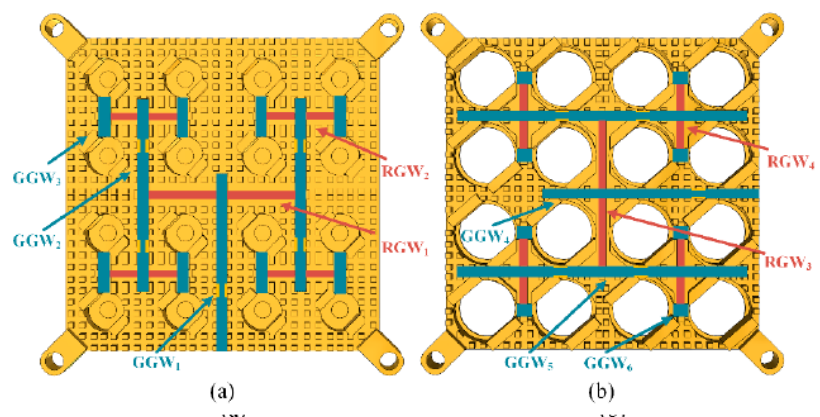


Figure 6. Top view of the horizontal cut planes going through two different layers of the proposed array. (a) Higher-frequency layer. (b) Lower-frequency layer. (GGWs: blue; RGWs: red.).

Figure 7a shows the topology of the RGW-to-GGW power dividers used in the feed network. The protrusion at the end of the RGW can be used to optimize the input reflection coefficient of the power divider. Figure 7b shows the topology of the GGW-to-RGW dividers. The distance between the RGWs and the shorter end of the GGW is around $\lambda_g/4$, where λ_g is the wavelength of the guided wave in the GGW. This distance is chosen to generate the maximum E-field value at the crossing point between the GGW and RGWs to excite the parallel RGWs. It should be noticed that the RGWs have no protrusion in a GGW-to-RGW power divider structure. The dimensions of the RGW-GGW transitions are summarized in Tables 1 and 2.

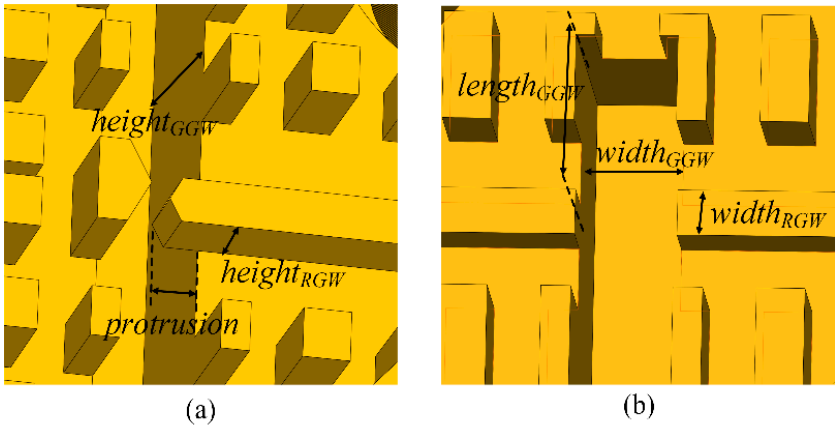


Figure 7. (a) RGW-to-GGW divider. (b) GGW-to-RGW divider. The fundamental parameters of each waveguide are indicated.

Figure 8 shows the electrical field distribution of a section of the feed network that contains both GGW-to-RGW and RGW-to-GGW transitions. The out-of-phase field distribution in the two RGW sections to the right and to the left of the crossing point with the GGW can be clearly observed. To compensate for the 180° phase difference without having to use any additional phase shift lines, the output GGWs are placed back to back, which cancels the 180° phase difference. As a result, sixteen radiation units are uniformly excited with the same amplitude and phase.

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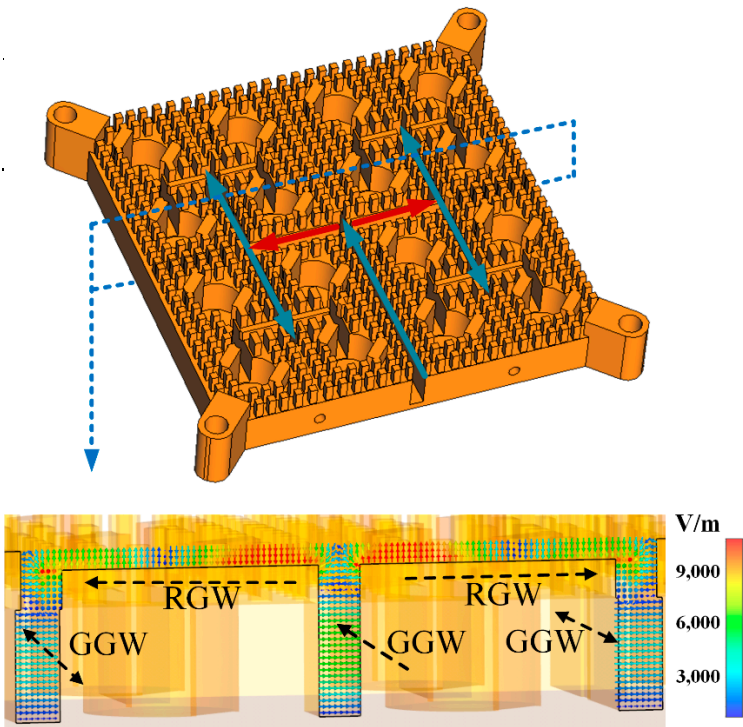
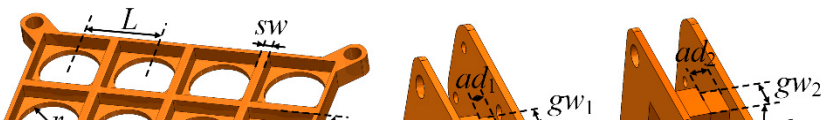


Figure 8. Electrical field representation between RGW and GGW.

Table 1. Dimensions of the GGW and the composite feed network (unit: mm).

	GGW ₁	GGW ₂	GGW ₃	GGW ₄	GGW ₅	GGW ₆
width	1.78	1.9	1.9	1.2	0.8	2.25
height	4.62	4.62	4.62	4.8	4.8	4.9
length	9.9	9.9	9.9	1.6	1.6	NA

Owing to the large distance ($L = 13\text{ mm}$, about $1.2\lambda_0$ at 28 GHz) between adjacent antenna units, grating lobes occur at Ka-band with high levels. To reduce the grating lobe levels, a waffle grid, whose structure is shown in Figure 9a, was designed to increase the effective aperture area of the array unit cell. More details about the mechanism of the waffle grid to mitigate grating lobes can be found in [19].



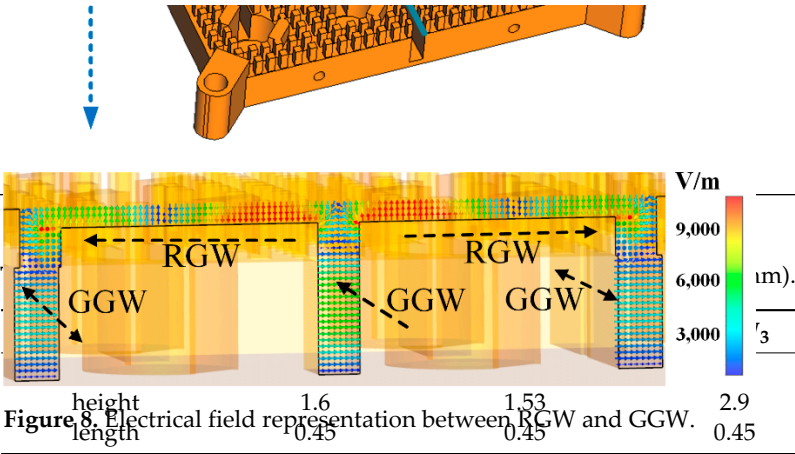


Figure 8. Electrical field representation between RGW and GGW.

		RGW ₄	
height	1.17	2.82	0.45
length	0.45	0.45	

2.4. Design of the Waffle Grid and Waveguide Transitions
Owing to the large distance ($L = 13$ mm, about $1.2\lambda_0$ at 28 GHz) between adjacent antenna units, grating lobes occur at Ka-band with high levels. To reduce the grating lobe levels, a waffle grid, whose structure is shown in Figure 9a, was designed to increase the effective aperture area of the array unit cell. More details about the mechanism of the waffle grid to mitigate grating lobes can be found in [19].

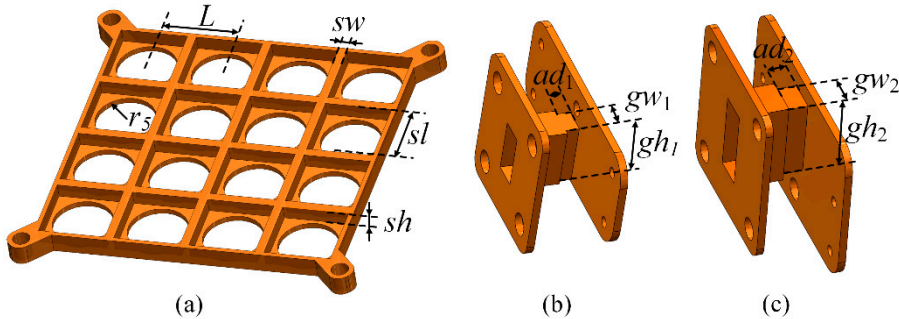


Figure 9. Grid layer and input-port transitions of the proposed antenna. (a) Waffle-grid layer. (b) Ka-band transition. (c) K-band transition. The parameters of the waffle grid and waveguide transitions follow: $L = 5.5$ mm, $SW = 1.5$ mm, $r_5 = 2.5$ mm, $sl = 1.5$ mm, $sh = 1.8$ mm, $ad_1 = 1.8$ mm, $gw_1 = 1.36$ mm, $gh_1 = 1.28$ mm, $gh_2 = 1.28$ mm, $ad_2 = 1.8$ mm, $gw_2 = 1.36$ mm, $gh_1 = 1.28$ mm, $gh_2 = 1.28$ mm.

To facilitate the connection of the proposed array with testing equipment, two waveguide stepped impedance transitions were designed to connect the respective inputs of the Ka- and K-band arrays to standard WR-28 and WR-42 waveguide ports. The Ka- and K-band transition structures are shown in Figure 9b,c, respectively. For each structure, the thickness of the transition wall is 0.5 mm.

3. Experimental Results and Discussions

3.1. Design of the Waffle Grid and Waveguide Transitions

A proprietary industrial-grade SLA 3D printer developed by the National Institute of Additive Manufacturing of China was used to fabricate the proposed antenna array. The resolution of the prototyping process is 50 μ m and the printing tolerance is ± 0.1 mm, which provides sufficient accuracy for the intended application. To metalize the 3D-printed prototype, a 10 μ m thick layer of copper was then electroless plated onto the resin, forming a conductive layer. The thickness of the copper layer is more than 20 times the skin depths at Ka/K-bands (about 0.47 μ m for copper at 20 GHz). Hence, the proposed antenna has the same electromagnetic properties as an all-metal antenna. It should be noted that the effective electrical conductivity of the plated copper layer is smaller than the typical value of 5.8×10^7 S/m for copper, according to [30]. Therefore, we used an effective conductivity of 1.5×10^7 S/m reported in this reference to model the plated copper in EM simulations of the antenna.

Photographs showing the fabricated prototype at different stages throughout the fabrication process are shown in Figure 10. Figure 10a shows the initial 3D-printed versions of the antenna array and waveguide transitions, which were then metalized, as shown in Figure 10b, and polished, as shown in Figure 10c. The unassembled antenna components are shown in Figure 10d. The complete antenna structure that was assembled using four M3 screws and had two standard waveguide ports mounted to it is shown in Figure 10e. Owing

to the contactless feature of GWs, assembling different layers of the antenna prototype only required bolting. With the SLA 3D printing technology, the average cost of each prototype is less than USD 50, which is about ten times cheaper than using the traditional CNC process. Each complete antenna structure weighs 35 g.

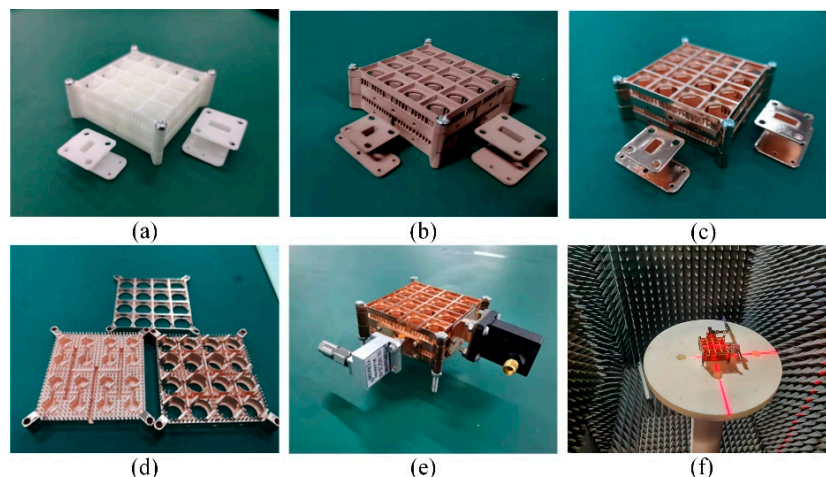


Figure 10. Fabricated antenna array prototype. (a) SLA-printed prototype. (b) Surface-metalized prototype. (c) Surface-polished prototype. (d) Unassembled parts. (e) Fully assembled antenna structure. (f) Measurement setup.

Figure 11 illustrates a sample of some crucial parts of the prototype array. Surface roughness due to the SLA printing and metallization are visible as ripples that can be seen in several places (see blue circles in Figure 11). However, since the surface of the prototype was polished, such ripples are merely 1 μm thick, and their effect is negligible. In addition, the right-angle corners of metallic pins and RGWs became rounded corners when fabricated (see purple circles in Figure 11), which have also been taken into account in the simulated model. Capacitive slits placed in a GGW for improving the impedance matching can be seen in Figure 11c. Because of fabrication error in the vertical dimension, the metallic pin array in Section 2, the air gap between the pins and the top metallic plate is intended to be 0.1 mm. Since the gap between the pins is less than a quarter of plate wavelength, the radiation from this air gap is not affected by the EBG characteristics. Even if, slightly, variations on the air gap do not affect the EBG characteristics. In fact, the gap pins are in direct contact with the top metallic plate (see Fig. 11d) with the gap used to guide the geometry of the antenna. The antenna is a planar antenna fed by a gap feed with the gap used to guide the geometry of the antenna. The antenna is a planar antenna fed by a gap feed with the gap used to guide the geometry of the antenna. The antenna is a planar antenna fed by a gap feed with the gap used to guide the geometry of the antenna.

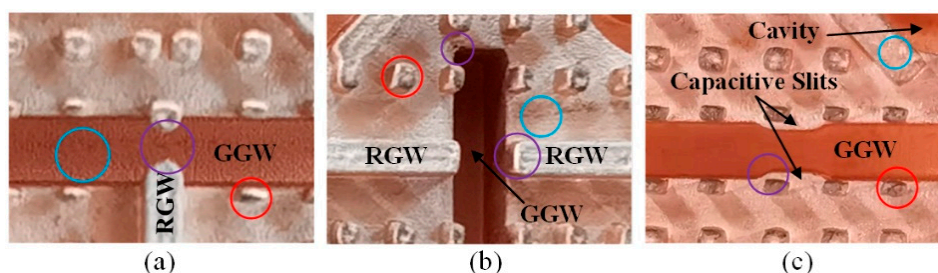


Figure 11: Photographs of some crucial parts of the antenna prototype: (a) RGW-to-GGW divider, (b) GGW-to-RGW divider, (c) Capacitive slots of GGW.

3.2. Experimental Results

The S-parameters of the fabricated array were measured by using an Agilent vector network analyzer (E8363B). Figure 12 shows the measured and simulated reflection coefficients of the proposed antenna. The measured input reflection coefficients are below -10 dB within the bandwidths 19.91–20.72 GHz (3.9%) in K-band and 28.22–29.19 GHz (3.5%) in Ka-band. The corresponding impedance matching bandwidths acquired from simulation results are 19.76–20.77 GHz and 27.45–29.41 GHz for the K- and Ka-bands, respectively. While the measured bandwidths are smaller than the simulated values, both meas-

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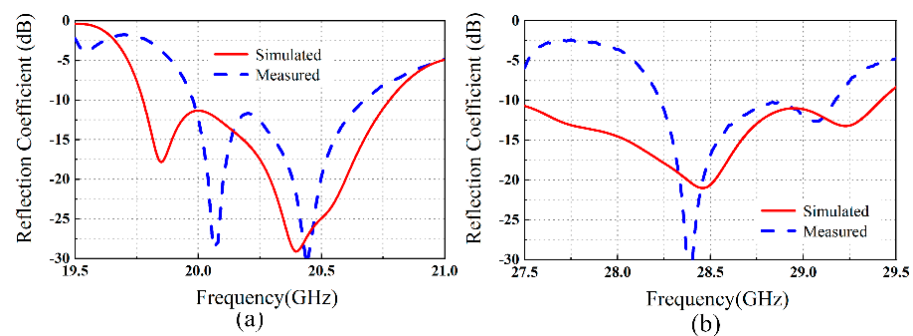


Figure 12. Simulated and measured reflection coefficient versus frequency. (a) S_{11} for lower frequency band. (b) S_{11} for higher frequency band.

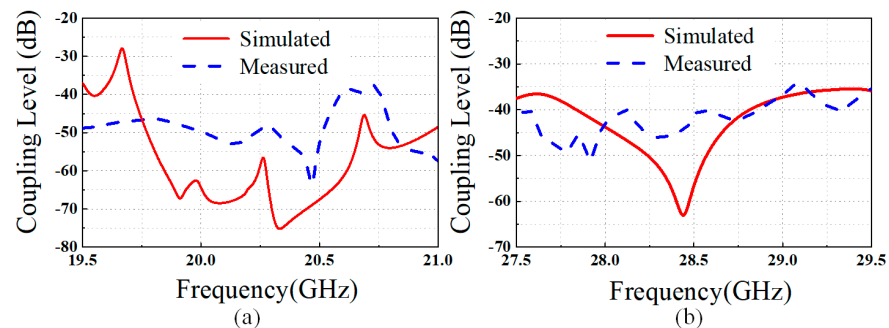
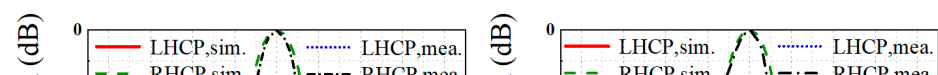


Figure 13. Simulated and measured inter-port coupling level versus frequency. (a) S_{12} for lower frequency band. (b) S_{21} for higher frequency band.

The far-field radiation patterns of the antenna were measured in an anechoic chamber, as shown in Figure 10f. Figures 14 and 15 present the simulated and measured normalized radiation patterns of the array in the x-z and y-z planes at 20 and 28.5 GHz. Overall, the measurement results are in good agreement with the simulation results. In the x-z plane, the 3 dB beamwidths of the RHCP and LHCP beams are 14° and 13° , respectively, and the sidelobe levels are lower than -13.3 and -12.2 dB, respectively. The patterns in the y-z plane are similar to those obtained for the x-z plane. Figure 16 shows the simulated and measured axial ratios (ARs) as functions of frequency over the lower and higher operating bands. The 3 dB AR bandwidths determined from the measurements are $19.89\text{--}20.52\text{ GHz}$ (3.1%) and $28.26\text{--}29.31\text{ GHz}$ (3.6%), which mostly overlap with the -10 dB impedance matching bandwidths for both bands. A noticeable difference can be seen between the simulated and measured results in a small fraction of the higher band for frequencies below 28.5 GHz. This can be caused by multiple factors, including fabrication error and measurement uncertainty of the far-field characterization system.



the x - z plane, the simulated and measured axial ratios are 21.1 and 20.7 dB, respectively, and the sidelobe levels are lower than -13.3 and -12.2 dB, respectively. The patterns in the y - z plane are similar to those obtained for the x - z plane. Figure 16 shows the simulated and measured axial ratios (ARs) as functions of frequency over the lower and higher operating bands. The 3 dB AR bandwidths determined from the measurements are 19.89–20.52 GHz (3.1%) and 28.26–29.31 GHz (3.6%), which mostly overlap with the -10 dB impedance matching bandwidths for both bands. A noticeable difference can be seen between the simulated and measured results in a small fraction of the higher band below 28.5 GHz. This can be caused by multiple factors, including fabrication error and measurement uncertainty of the far-field characterization system.

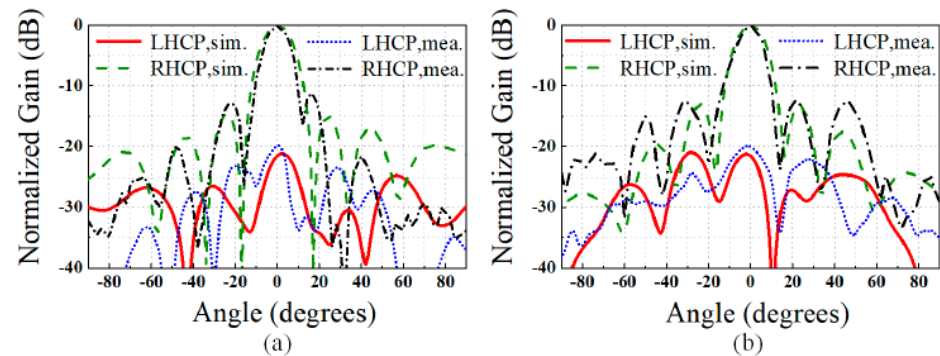


Figure 14. Simulated and measured normalized radiation patterns at 20 GHz in the (a) x - z plane and (b) y - z plane.

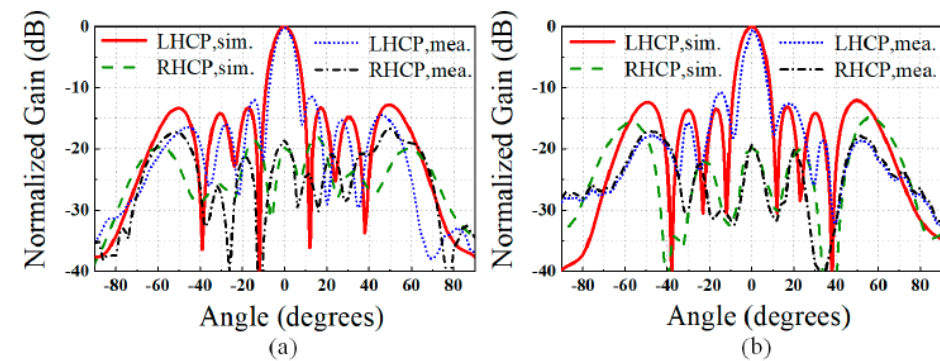


Figure 15. Simulated and measured normalized radiation patterns at 28.5 GHz in the (a) x - z plane and (b) y - z plane.

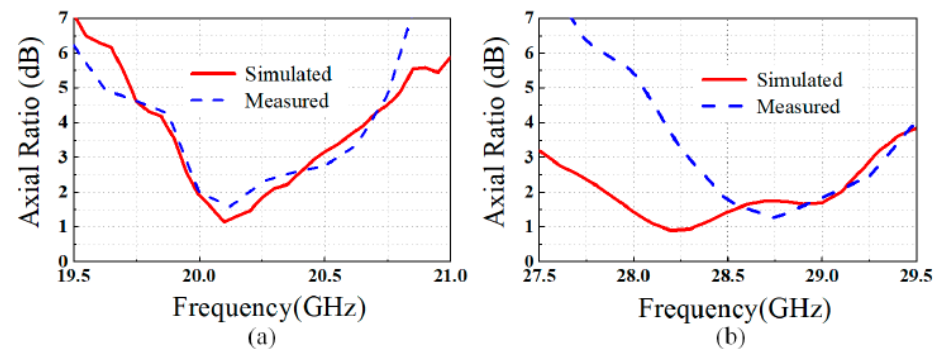


Figure 16. Simulated and measured axial ratio for the (a) lower frequency band and (b) higher frequency band.

The gains and simulated efficiency of the antenna array at several frequency points within its two operating bands are summarized in Table 3. The maximum measured gains are 20.1 and 20.3 dB for the lower and higher band, respectively. The antenna prototype provides efficiency of at least 88.9% over both operating frequency bands. Given the aperture size 52 mm $\times$ 52 mm, the calculated aperture efficiencies of the array are 66.25% at 20 GHz and 32.6% at 28.5 GHz. The results indicate that the proposed dual-band, dual-CP antenna can simultaneously achieve high efficiencies, high aperture efficiencies, and high gains.

Table 3. Gain and efficiency of the proposed antenna.

Frequency (GHz)	20.25	20.5	20.75	28	28.5	29
Polarization	RHCP	RHCP	RHCP	LHCP	LHCP	LHCP
Sim. gain (dB)	21.1	21.1	20.8	20.5	20.8	21.2
Mea. gain (dB)	20.0	20.1	19.8	19.9	20.3	20.2
Efficiency (%)	92.1	91.5	88.9	91.3	91.5	89.1

To highlight the advantages of the proposed antenna, we compared the proposed antenna with other similar designs and summarized the comparison in Table 4. The dual-band, dual-CP array reported in [7] has a wide AR bandwidth due to the sequentially rotated feed network. However, the higher- and lower-frequency elements of this array are coplanar, resulting in a significantly larger electrical aperture size than a stacked structure design similar to our proposed approach. Compared with other dual-band, dual-CP arrays implemented using PCB lithography in [8,9], the proposed array has a higher efficiency and power capacity due to the full metallic-body construction that helps avoid dielectric losses. Moreover, the combination of dual-band, dual-CP capability with high efficiency, compact aperture size, and low-cost fabrication emphasizes the significance of our proposed design in comparison to various GW-based antenna works [19–22]. Indeed, the antenna presented in the current work provides the highest efficiencies among all GW-based, dual-band, dual-polarization antennas listed in Table 4. The only work reporting a slightly higher efficiency than our current study is [19], which presented a single-band, single-CP antenna. Furthermore, the SLA 3D printing process significantly reduces the cost, fabrication complexity, and weight of the antenna structure when compared to using the conventional CNC machining method.

Table 4. Comparison with some relevant reported work.

Reference	Band	Polarization	Gain (dB)	Efficiency (%)	Technology	Aperture Size ($\lambda_L \times \lambda_L$)
[7]	X/Ku	Dual-CP	17.5/18.3	–	PCB	28.4×25.6
[8]	K/Ka	Dual-CP	13.2/11.5	78/87	PCB and GW	1.53×1.53
[9]	X/Ku	Dual-CP	16.1/15.2	–	PCB	2.9×1.7
[19]	Ka	Single-CP	22.0	98	CNC and GW	5.12×5.12
[20]	Ka	Dual-LP	27.5	75	CNC and GW	10×10
[21]	Ka	Dual-CP	27.0	75	CNC and GW	9.7×9.7
[22]	K/Ka	Dual-LP	25.8/29.5	84/88	CNC and GW	6.3×6.3
This work	K/Ka	Dual-CP	20.1/20.3	92.1/91.5	SLA and GW	2.6×2.6

λ_L denotes the free-space wavelength referring to the lowest frequency of the corresponding antenna; LP denotes linear polarization.

4. Conclusions

In this paper, a 3D-printed, GW-based, dual-band, dual-CP antenna array is presented. The K-band array is stacked above the Ka-band array to share the antenna aperture, resulting in an efficiently packaged, compact antenna structure. The RHCP and LHCP waves are excited in cylindrical cavities with chamfered edges. The use of GW not only results in high antenna efficiency at the K- and Ka-bands, but also decreases the sensitivity of the antenna performance to fabrication tolerances, which allowed us to leverage low-cost prototyping techniques. The antenna prototype consisting of 4×4 radiating elements was fabricated using SLA 3D printing technology in conjunction with metallization and surface polishing processes, which greatly decreases the cost and weight compared to conventional CNC technology. The fabricated prototype exhibited gains of about 20 dB in both operating bands. Moreover, the design architecture can be seamlessly scaled to build a larger array without needing any additional layers. Owing to its high gain, high efficiency, scalability, low cost and lightweight characteristics, the proposed antenna design is a potential solution for commercial millimeter wave communication.

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