A high-gain circular polarization beam scanning transmit array antenna

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Abstract
A high-gain circular polarization (CP) beam scanning antenna based on planar transmit array (TA) is proposed. The planar TA is implemented using rectangular patches and square rings, and it has different frequency responses to two orthogonal polarized incident waves owing to different phase distributions in x- and y-direction. The TA can convert a linearly polarized spherical incident wave into a circularly polarized transmitted plane wave. In addition, beam scanning is achieved by in-plane translation of the feed antenna. The procedure is demonstrated for high-gain beam steering implementation operating in CP at the Ku-band. Full-wave simulation results demonstrate that the designed antenna enables elevation beam scanning from −33.6° to +33.6°, with a peak gain of 26.4 dBi. Scanning loss is better than 5 dB and the axial ratio remains below 3 dB for the entire scanning range in the frequency band of 13.9–14.9 GHz. Moreover, a prototype is fabricated and the measured results have good agreement with the simulated results, which confirms the feasibility of using this technology to achieve a high-gain, CP, low-cost and low cross-polarization beam scanning antenna.

INTRODUCTION
In modern wireless communications, the research and design of an antenna that has high-gain, circular polarization (CP), and beam scanning performance while maintaining low cost, a low profile, and ease of fabrication has important practical significance. CP waves have the ability to restrain interference from rain and fog, suppressing multipath effects, and so on [1]. High-gain and beam scanning performance are required for the link budget and to maintain links on-the-move [2]. A low cost and easy fabrication are needed for wide application. Traditional CP phased array antennas can achieve beam scanning, but they have some obvious drawbacks such as a large insertion loss of the feed network and the use of many expensive transceiver modules [3]. Moreover, the axial ratio will be degraded when the scan angle is moved away from broadside [4].

In addition to circularly polarized phased array antennas, there are many other techniques for achieving CP beamsteering, such as a mechanical CP array [5, 6], a pattern reconfigurable CP antenna [7–9], a CP antenna array based on the Butler matrix [4, 10, 11] and a transmit array (TA) antenna [2, 12]. Mechanical CP arrays tend to be bulky and inefficient to fabricate at high frequencies. Pattern reconfigurable CP antennas and CP antenna arrays based on the Butler matrix cannot achieve continuous beam scanning and have complex structures. The TA antenna, which consists of a flat surface illuminated by a feed antenna, is a promising alternative solution.

In the reported literature, most research on CP TAs are limited to fixed beam and polarization conversion techniques such as element sequential rotation have been widely used in designs [13–19]. Some works can achieve CP beam steering based on TAs [2, 20–23]. In Lima et al. [2], an appropriately designed thin flat lens is translated horizontally over a fixed CP primary feed, providing a wide elevation steering range without the need for lens tilting. The work reported in Hasani et al. [22] is a novel dual-band polarization-independent TA fed by a dual-band circularly polarized ridged cavity antenna. A dual-band wide-angle CP beam steering TA antenna with a novel and low-loss dual-band dual-LP-to-CP converter is presented in Naseri et al. [23]. In these CP beam scanning designs, the generation of CP waves and the realization of beam scanning are two separated processes. The proposed TAs are generally composed of isotropic elements and have the same frequency...
responses to two orthogonal polarized incident waves. Horn antennas combined with orthomode transducers, CP patch antennas, and linear polarization (LP) antennas with polarization converters are often used as the feed antenna. However, these circularly polarized feed solutions are either bulky and expensive or inefficient and hard to fabricate up to millimetre-wave frequencies [23].

A method for designing high-gain CP beam scanning TA antennas with LP feeds is presented. The proposed antenna is composed of an LP horn antenna and a planar lens. The lens can convert LP spherical incident waves into a CP transmitted plane wave, and beam scanning is achieved by in-plane translation of the planar lens or the feed antenna. The innovation of the proposed design is that polarization conversion and beam scanning are simultaneously achieved with the same TA while exhibiting high-gain and low cross-polarization. Moreover, as the beam deflection angle increases, the axial ratio can always be kept below 3 dB and does not deteriorate. This method reduces the design complexity and the cost of CP beam scanning antennas. The procedure is demonstrated at the Ku-band. The simulation and experimental results prove the feasibility of using this technology to achieve a high-gain, CP, low-cost and low cross-polarization beam scanning antenna.

This work is organised as follows. Section 2 elaborates on the design principle. Section 3 presents the unit of TA and the configuration of the entire structure. In Section 4, the simulation and measured results of the proposed CP beam scanning antenna are presented. Finally, conclusions are drawn in Section 5.

2 | OPERATION PRINCIPLES OF THE DESIGNED TA

The schematic representation of the CP beam scanning antenna is presented in Figure 1. A planar lens in the \( z = 0 \) plane is fed by an LP standard horn antenna located at \( (0, 0, -F) \). The origin of the coordinate system is fixed at the centre of the lens. The electric field vector of the radiation LP waves is tilted \( \pi/4 \) relative to the \( x \) and \( y \) axes, and it has the same electric field components, \( E_x, E_y \), in the two orthogonal directions. When a linearly polarized spherical wave is incident perpendicular to the surface of the planar lens, the transmitted wave will be converted into a circularly polarized plane wave owing to different phase distribution in the \( x \) and \( y \) direction. Moreover, beam scanning in the \( xoz \) plane can be achieved by changing the relative position of the lens and the feed. The change of relative position can be realized either by individually moving the feed antenna or simply by translating the lens along the \( x \) axis. The next section explains the implementation of polarization conversion and beam scanning.

2.1 | Implementation of polarization conversion

In Figure 1, when the feed antenna of the planar lens is located at \( (a, 0, -F) \), the spherical phase distribution \( \phi_{\text{in}}^{x,y} \),

\[
\phi_{\text{in}}^{x,y}(x,y) = \phi_{\text{in}}^{x}(x,y) - \phi_{\text{in}}^{y}(x,y)
\]

\[
= \frac{2\pi f}{c} \sqrt{(x-a)^2 + y^2 + F^2} - \phi_{x,y}^{\text{in}}
\]

\[
\phi_{\text{in}}^{x}(x,y) - \phi_{\text{in}}^{y}(x,y) = \zeta_y - \zeta_x
\]

\[
\phi_{\text{in}}^{y}(x,y) = \frac{2\pi f}{c} \sqrt{(x-a)^2 + y^2 + F^2} + \xi_x
\]

\[
\phi_{\text{in}}^{x}(x,y) = \frac{2\pi f}{c} \sqrt{x^2 + y^2 + F^2} + \zeta_x
\]

\[
\phi_{\text{in}}^{y}(x,y) = \frac{2\pi f}{c} \sqrt{x^2 + y^2 + F^2} + \zeta_y
\]

\[
\phi_{\text{in}}^{x}(x,y) = \phi_{\text{in}}^{x,y}(x,y) + \zeta_x
\]

\[
\phi_{\text{in}}^{y}(x,y) = \phi_{\text{in}}^{x,y}(x,y) + \zeta_y
\]
Equation (5) presents the transmission phase difference in two orthogonal directions. If $\zeta_x - \zeta_y = \pm \pi/2$ and the planar lens has the same transmission amplitude response for the two orthogonal components of the electric field, the linearly polarized incident wave will be successfully converted into a circularly polarized wave.

2.2 | Implementation of beam scanning

As analysed in Lima et al. [2], to understand the relation between the feed position and the beam scanning angle, a Taylor expansion of Equation (4) around the point $(x = a/2, y = 0)$ is performed. It can be expressed as a sum of a linear term $\varphi_i^{x,y}$ with all remaining non-linear terms $\varphi_{i\alpha}^{x,y}$:

$$
\varphi_i^{x,y}(x,y) = \varphi_i^{x,y}(x,y) + \varphi_{i\alpha}^{x,y}(x,y)
= \zeta_{x,y} + \frac{2\pi f}{c} \frac{a}{\sqrt{a^2 + 4F^2}}
- \frac{2\pi f}{c} \frac{2a}{\sqrt{a^2 + 4F^2}}
+ \varphi_{i\alpha}^{x,y}(x,y)
$$

(6)

Figure 2 shows the phase distribution curves of the transmitted waves along the $x$ axis. The solid curves represent the theoretical transmission phase distributions $\varphi_i$ for different feed locations $(a, 0, -F)$. The dashed curves represent the phase values corresponding to linear term $\varphi_i$. The differences among them represent phase values corresponding to the non-linear term $\varphi_{i\alpha}$. As shown in the figure, when the feed is located at $(a, 0, -F)$, within a certain region of the lens, linear term $\varphi_i^{x,y}$ can approximate function $\varphi_{i\alpha}^{x,y}$. In that region, $\varphi_i^{x,y}$ can be used reasonably to replace the $\varphi_{i\alpha}^{x,y}$. Transmission phases $\varphi_i^{x,y}$, $\varphi_i^{y,z}$ of the two orthogonal components of the electric field can be expressed by Equations (7) and (8):

$$
\varphi_i^{x,y}(x,y) = \zeta_x + \frac{2\pi f}{c} \frac{a^2}{\sqrt{a^2 + 4F^2}}
- \frac{2\pi f}{c} \frac{2a}{\sqrt{a^2 + 4F^2}}
$$

(7)

$$
\varphi_i^{y,z}(x,y) = \zeta_y + \frac{2\pi f}{c} \frac{a^2}{\sqrt{a^2 + 4F^2}}
- \frac{2\pi f}{c} \frac{2a}{\sqrt{a^2 + 4F^2}}
$$

(8)

The transmission phase of the two components of the electric field has the same linear gradient in the $x$ direction. Thus, the beams of two orthogonally polarized waves can be deflected into the same direction. Beam deflection angle $\theta(a)$ can be estimated from Equations (7) and (8) as

$$
\theta(a) = -\arcsin\left(\frac{2a}{\sqrt{a^2 + 4F^2}}\right)
$$

(9)

Once feed position $a$ and focal length $F$ are determined, beam deflection angle $\theta$ can be calculated by Equation (9). Moreover, when $a > 0$, $\theta(a)$ is a negative value. The beam will be pointed to the direction of $\theta = 180^\circ$. When $a < 0$, $\theta(a)$ is a positive value. The beam will be pointed to the direction of $\theta = 0^\circ$. Moreover, as can be seen in Figure 2, with an increase in the feed offset distance, the region where $\varphi_i^{x,y} = \varphi_{i\alpha}^{x,y}$ decreases and the non-linear phase error effects increase. This would lead to an increase in beam pointing angle error.

3 | STRUCTURE OF CIRCULAR POLARIZATION BEAM SCANNING ANTENNA

3.1 | Unit cell design

In the lens unit design, it is necessary to ensure that the transmission amplitudes of the units are as high as possible and the transmission phase variation extends over a range of $[0, 360^\circ]$. High transmission amplitudes are required to guarantee high efficiency of the lens and a $360^\circ$ phase shift is needed to achieve full control of the transmission wavefront. In the design, phase variation is realized by changing the unit size. In general, increasing the number of layers of the unit can effectively expand the maximum achievable phase shift range. However, with an increase in the number of layers, the loss of the dielectric layers and the thickness and weight of the structure will increase, and the lens efficiency will decrease. Therefore, in the case in which the range of transmission phase variation satisfies the requirements, the number of layers should be as small as possible.

Figure 3 shows the basic structure of the planar lens unit, which is similar to that reported in Cai et al. [24]. It is composed of four identical metal layers and three intermediate dielectric layers. Figure 3(a) shows the top view of the metal layers. Each metal layer contains a square ring outside and a
rectangular patch inside. The dimension of the unit cell is $p = 6.4 \text{ mm}$, the length of the square microstrip ring is $w = 0.12 \text{ mm}$, and the length of the inner rectangular patch in the $x$ and $y$ directions are $L_x$ and $L_y$, respectively. The transmission phase change is achieved by adjusting the length of the inner patch, $L_x$ and $L_y$. The metal layers are separated by three dielectric layers. Each dielectric layer has a thickness of $h = 1.5 \text{ mm}$ and dielectric constant of 2.65. The design and optimization of the proposed unit are performed in CST Microwave Studio. To simulate an infinite periodic structure, unit cell boundary conditions are assigned along the $x$ and $y$ directions and open boundary conditions are applied in the $z$ direction. Moreover, the units are illuminated by two normal plane waves propagating in the $z$ direction with the electric fields along the $x$ and $y$ directions.

Figure 4 presents the transmission phase and amplitude of the proposed lens unit with different sizes of the inner patch at 14.2 GHz. To ensure that the structure has the same frequency response for $x$- and $y$-polarized normal incident waves, $L_x = L_y$ is set; that is, each metal layer has a square patch inside. As shown in Figure 4, with an increase in the length of patch $L_x$ from 0.5 to 5.55 mm, the transmission phase varies from $-210^\circ$ to $-662^\circ$. In addition, the transmission amplitudes are all greater than 0.84. To explain the reason for designing such a unit structure with three dielectric layers and four metal layers, two other different unit structures are also characterized using CST software. Figure 5(a) and (b) shows the simulated results for the unit with a single layer and two dielectric layers, respectively. Under the condition of transmission amplitudes greater than 0.8, the maximum achievable phase shift range is about $187^\circ$ for the unit with a single dielectric layer and $322^\circ$ for the one with two dielectric layers. Using these two units cannot achieve $[0, 360^\circ]$ phase range. The three dielectric layers unit is the one with the least number of layers that can satisfy the requirements of phase change with identical metal layers, so it was used for designing the planar lens.

In addition, the proposed unit is polarization-independent, which is similar to that reported in Xu et al. [25]. To illustrate it, parameter $L_x$ is kept constant and a parameter sweep analysis of $L_y$ is performed. As shown in Figure 6, the transmission phase decreases as $L_y$ changes from 1 to 5.4 mm and the maximum phase change is close to $360^\circ$ for $y$-polarized waves. Also, the transmission phase variation is less than $19^\circ$ for $x$-polarized waves. A comparison of the frequency response of two linear polarization waves shows that the proposed unit is polarization-independent. Therefore, the transmission phase of the $x$- and $y$-polarized waves can be controlled independently by adjusting the values of $L_x$ and $L_y$. Thus, the wavefront and polarization of the transmission waves can be controlled effectively.

### 3.2 Implementation of planar lens antenna

Through Equations (2) and (3), the continuous phase distributions of the planar lens in the $x$ and $y$ directions can be obtained with $\zeta_x = 0, \zeta_y = \pi/2$. Because $\zeta_y - \zeta_x = \pi/2$, the transmission wave will be right-hand CP in Equation (5). In the design, a series of discrete phase elements are used to represent the phase distribution of the designed lens. It is known that $p$ is the unit period; thus the Equations (2) and (3) can be rewritten as

$$\phi_{\text{lin}}^x(x, y) = \frac{2\pi f}{c} \sqrt{(mp)^2 + (np)^2 + F^2} \tag{10}$$

$$\phi_{\text{lin}}^y(x, y) = \frac{2\pi f}{c} \sqrt{(mp)^2 + (np)^2 + F^2 + \frac{\pi}{2}} \tag{11}$$
According accurately and for the calculated responding As for software and in relation of dimensions according unit the n. Microw of unit each m. um. achieving 2 getting has and focal av ty the onal obtain, and maxim at can the of as where ϵm index aper d diameter 14.2 phase Studio directivi lense in its CST h used in of the ulae, is m to hievable a The represents x um physical (planar trans-the lens x ortho according of next e length simulate tw and and antenna Ly size al. mance definite analysed units angle ac Z ty be of y estimate cor size. a these is to to F (GHz. the to maxim the transmission transmission lens wa proposed orthog. polarized phase fixe unit (b) ch to ves FIGURE 6 Lx (a) o T danging for different Single of w transmission dielectric (a) o layers and of phase units T (Z) Lx (a) to T danging for the proposed lens unit with fixed Lx and changing Ly. (a) y polarisations; (b) x polarisations

where m, n ∈ Z and m (n) represents the index of the unit in the x(y) direction. According to these formulae, the transmission phase of each unit in two orthogonal directions can be accurately calculated with a definite focal length F at 14.2 GHz. According to the transmission phases of each unit in x and y directions, CST Microwave Studio software is used to simulate and obtain corresponding physical dimensions Lx and Ly.

The next step is to estimate the size of the planar lens and the number of units according to the required antenna performance such as the maximum achievable directivity and the maximum beam deflection angle. The maximum achievable directivity of the lens antenna has a close relation to its aperture size. As analysed in Lima et al. [2], aperture diameter $D_\alpha$ is determined by prescribed directivity $D_{\text{max}}$ and edge field taper level $10^{-x}$, and it can be calculated by

$$D_\alpha = \frac{\lambda}{4\pi} \sqrt{\frac{8\pi \ln 10}{\tanh(\tau/2 \ln 10) \cos a_0 \eta} D_{\text{max}}}$$

(12)

where $\eta$ defines the aperture efficiency and is the multiplication of transmission efficiency of the planar lens, aperture-illumination efficiency, and spillover efficiency. $a_0$ represents the beam deflection angle from the broadside direction. If the directivity of a planar lens antenna required is 26 dBi at $a_0 = 10^o$, with $\eta = 50%$ and $\tau = 0.5$ corresponding to a $-10$ dB edge taper, aperture diameter $D_\alpha$ can be calculated to be 195.6 mm by Equation (12). Thus, the whole structure will be composed of 31 × 31 units if the structure units are arranged in a square, corresponding to an aperture size of 198.4 × 198.4 mm. In addition, $F = 120$ mm is chosen here and $F/D_\alpha \approx 0.6$.

Furthermore, to achieve beam scanning in the xoz plane, the feed antenna needs to be translated in the x direction. The dimension of the lens in the x direction should be larger than that in the y direction, although this will reduce the aperture efficiency. The additional size $D_\alpha$ can be calculated according to the maximum beam deflection angle $\theta_{\text{max}}$ to be achieved. If $\theta_{\text{max}} = 30^o$, $D_{\text{in}} \approx 62.0$ mm will be calculated using Equation (9). Because the increase in the structure size can only be an integral multiple of unit period $p$, 10 more units are added in the ±x directions.

Finally, the aperture size is $326.4 \times 198.4$ mm with 51 units in the x direction and 31 units in the y direction. The
focal length of the planar lens is 120 mm. Based on Equations (10) and (11), the dimensions of each unit can be obtained by parameter optimization. Figure 7 shows the top view of the designed planar lens, which consists of 1581 units. The feed antenna is a linearly polarized standard horn antenna; its polarization direction is tilted 45° with respect to the x axis and y axis to ensure the same electric field component in both directions for L-C polarization conversion. To demonstrate the performance of the feed antenna, the simulated S11 and realized gain, and the E- and H-plane patterns at 14.2 GHz are presented in Figure 8(a) and (b), respectively.

4 | SIMULATION AND MEASUREMENT RESULTS

Based on this theoretical analysis and design, the entire model including the planar lens and the feed antenna is simulated in CST Microwave Studio. The position of the planar lens remains fixed and the centre position of its bottom surface is defined as the origin (0, 0, 0). The phase centre of the horn antenna is located at \((a, 0, -P)\). By changing the value of \(a\), the radiation patterns and axial ratio of the structure with different feed positions can be obtained. In the simulation, \(a\) is taken as a series of discrete values to illustrate the performance; they are all selected as integer multiples of lens unit period \(P\). Figure 9 presents the simulated 3D radiation patterns with different feed locations at 14.2 GHz. As the figure shows, with the change in feed position, the beam pointing direction is significantly deflected. The simulated far-field radiation patterns in the elevation plane are shown in Figure 10. According to the simulation results, the main beam of the lens antenna steers from \(+33.6°\) to \(-33.6°\), changing from \(-76.8°\) to \(+76.8°\). The realized gain drops and the beam widens with an increase in the scanning angle. Table 1 gives the accurate beam deflection angles and realized gains with different feed locations at 14.2 GHz. The realized gain in the broadside direction reaches a maximum of 26.4 dB, and the scan loss is about 5 dB. The transmission aperture efficiency is about 22% based on the maximum gain value.

![Top view of simulated planar lens](image)

**FIGURE 7** Top view of simulated planar lens

![Performance of linearly polarized feed antenna](image)

**FIGURE 8** Performance of linearly polarized feed antenna. (a) S11 and realized gain; (b) Simulated E-plane and H-plane radiation patterns at 14.2 GHz
The reason is that to increase the scanning angle, the sizes of the transmission array in the $\pm x$ axis direction are increased, which leads to a decrease in the efficiency of the whole array.

The axial ratio curves at different beam pointing directions are shown in Figure 11. Owing to the symmetry of the lens structure, only the axial ratio curves with negative $a$ are presented. At different feed locations, the axial ratios of the transmitted wave are all below 3 dB in the beam pointing directions. The simulated results confirm that the proposed method can achieve high-gain CP continuous beam scanning. Furthermore, the axial ratios versus frequency with different beam deflection angles are presented in Figure 12. As the main beam of the antenna is deflected from $0^\circ$ to $33.6^\circ$, the axial ratios in the main beam directions can always be kept below 3 dB in the frequency band of 13.9–14.9 GHz. Moreover,

Table 1 Comparison of simulated and measured results

| $a$ (mm) | Simulated beam tilt angle ($^\circ$) | Simulated realized gain (dB) | Measured beam tilt angle ($^\circ$) | Measured realized gain (dB) |
|---------|-----------------------------------|-----------------------------|-----------------------------------|----------------------------|
| 0       | 0                                  | 26.4                        | 0                                  | 26.1                       |
| -12.8   | 5.0                                | 26.0                        | 4.9                               | 25.8                       |
| -25.6   | 9.9                                | 25.4                        | 9.7                               | 25.0                       |
| -38.4   | 15.4                               | 24.3                        | 15.6                              | 24.1                       |
| -51.2   | 21.7                               | 23.5                        | 20.8                              | 23.4                       |
| -64.0   | 27.7                               | 22.5                        | 26.4                              | 22.3                       |
| -76.8   | 33.6                               | 21.4                        | 32.4                              | 21.0                       |

Figure 12 shows the realized gain versus frequency at different beam deflection angles. The realized gain decreases with the beam scanned away from the zenith. Table 1 shows the variation in realized gain at different deflection angles compared with the maximum gain in the broadside direction at 14.2 GHz.

To verify the simulated CP beam scanning performance, a prototype was fabricated using printed circuit board technology. The top view of the planar lens, which is composed of four metal layers and three dielectric layers, is shown in Figure 13(a). Its dimensions are $340 \times 210$ mm, which is slightly larger than that of the metal layer, for ease of assembly.

The total thickness is 6 mm. As shown in Figure 13(b), the lens antenna was measured with a frequency domain planar near-field test system in an anechoic chamber. A rectangular waveguide probe is used to test in the near-field area of the antenna to obtain the amplitude and phase of each sampling point. Then, the near-to-far field transformation programme is
used to get the far-field pattern of the antenna. Figure 14 presents the measured radiation patterns in the elevation plane with different positions of feed antenna.

The measurement results have good agreement with the simulated results. Their comparisons are presented in Table 1.

![Figure 11: Simulated axial ratios of lens antenna in xoz plane for different values of \( \alpha \) at 14.2 GHz](image)

![Figure 12: Simulated realized gain and axial ratios versus frequency at different beam tilt angles](image)

The data analysis of Table 1 reveals that the measured realized gain has a maximum difference of 0.4 dB from the simulated result. The discrepancies in the beam tilt angle are within 1.3° for all cases. Furthermore, the axial ratios of the lens antenna are measured with different feed locations; the measured results are presented in Figure 15. As can be seen from the figure, the axial ratios in the main beam direction can always be kept below 3 dB.

Although there are some small differences in the values of axial ratios, the performance of polarization conversion can be successfully verified. The measured axial ratios versus frequency at different beam tilt angles are presented in Figure 16. The measured results are closely agree with the simulated results in the frequency band of 13.9–14.9 GHz. The discrepancies could be attributed to undesired phase errors and assembly errors. Moreover, the simulated and measured realized gain and cross-polarization at 14.2 GHz for different beam tilt angles are presented in Figure 17. The cross-polarization is smaller than −18.5 dB for the entire scanning range.

A CP beam scanning TA antenna with an LP feed is designed. The design can be used for satellite communication, point-to-point mobile communications, and so on. Table 2 shows different design schemes of a CP beam scanning TA antenna in the references. A comparison shows that the proposed design is simpler and more flexible, which provides a highly efficient and inexpensive method for CP beam scanning design.

![Figure 13: Photograph of the whole structure, (a) Fabricated transmit array; (b) Measurement setup](image)

![Figure 14: Measured far-field radiation patterns in xoz plane with different feed locations at 14.2 GHz](image)
A method for designing a high-gain CP beam scanning antenna based on a TA was introduced. The proposed TA can cover a linearly polarized spherical incident wave into a CP transmitted plane wave. Moreover, beam scanning is realized simply by translating the feed antenna. The procedure is demonstrated at the Ku-band, and a TA is fabricated and measured to verify the design. The simulated and measured results are in good agreement. In the design, there is no need for TR components or a complex feed network. Beam deflection and polarization conversion are realized simultaneously. It reduces the demand on the feed antenna and the cost, with good performance. This method provides a new avenue for designing a high-gain 1D CP beam scanning antenna.

**TABLE 2** Comparison of one-dimensional circular polarization (CP) beam scanning design

| Reference | The implementation of 1D CP beam scanning |
|-----------|------------------------------------------|
| [2]       | CP patch antenna + transmit array (TA)    |
| [22]      | CP ridged cavity slot antenna + TA        |
| [23]      | LP horn antenna + polarization conversion metasurface + TA |

This work LP horn antenna + TA

Abbreviation: LP, linear polarization.

5 | CONCLUSION

A method for designing a high-gain CP beam scanning antenna based on a TA was introduced. The proposed TA can cover a linearly polarized spherical incident wave into a CP transmitted plane wave. Moreover, beam scanning is realized simply by translating the feed antenna. The procedure is demonstrated at the Ku-band, and a TA is fabricated and measured to verify the design. The simulated and measured results are in good agreement. In the design, there is no need for TR components or a complex feed network. Beam deflection and polarization conversion are realized simultaneously. It reduces the demand on the feed antenna and the cost, with good performance. This method provides a new avenue for designing a high-gain 1D CP beam scanning antenna.

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