Generation of Plane Spiral Orbital Angular Momentum Waves by Microstrip Yagi Antenna Array

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ABSTRACT

Compared with the orbital angular momentum (OAM) wave whose 3D radiation pattern is a central-hollowed cone, the plane spiral orbital angular momentum (PSOAM) wave propagates along the transverse direction, which is far more convenient for practical applications. In this paper, traditional microstrip Yagi antennas are formed into a uniform circular array (UCA) to radiate PSOAM electromagnetic waves, which has the outstanding advantage of a flat structure. Firstly, the fundamentals of generating PSOAM by microstrip Yagi antenna array is introduced. Secondly, a circular microstrip Yagi PSOAM antenna array fed by a self-designed Wilkinson feeding network is designed, fabricated and measured. Eight microstrip Yagi antenna elements and its feeding network are printed on the upper and lower substrate, respectively. The profile of the proposed antenna is only 0.16\(\lambda\), lower than other PSOAM antennas, which is convenient for integration with wireless communication systems. Both simulation and measurement results depict that it is capable of generating PSOAM with mode number \(l = 3\) on 3.43GHz. From 3.15GHz to 3.65GHz, the vortex electromagnetic wave of \(l = 3\) can be observed. The proposed microstrip Yagi antenna based on UCA provides a new way for the research of practical PSOAM antennas, has broad prospect in the future of the reconfigurable field.

INDEX TERMS

Microstrip Yagi antenna, plane spiral orbital angular momentum (PSOAM), uniform circular array (UCA), flat structure.

I. INTRODUCTION

As communication technology evolves rapidly, the contradiction between finite electromagnetic spectrum resources and continuously increasing bandwidth requirements has become increasingly prominent [1], [2]. As an inherent attribute of electromagnetic waves, orbital angular momentum (OAM) is expected to be a new degree of freedom that can be used to modulate and multiplex electromagnetic waves after amplitude, phase, frequency, and polarization [3], [4]. Most importantly, the OAM can theoretically provide infinite modes, and different modes are orthogonal [5], [6]. Therefore, the OAM may provide unlimited signal channels, expanding the capacity of the communication system and alleviating the spectrum tension instead of expanding bandwidth [1], [7], [8].

Additionally, owing to the electromagnetic wave carrying OAM has the characteristic of spiral phase wavefront, it has outstanding advantages in the field of detection and imaging [9]–[12], especially in the application of radar to the recognition of rotating targets [13]–[15]. Besides, its application in the Doppler detection [16], [17], and quantum communication which has high confidentiality [18] are also promising. However, the E-field radiation pattern of the OAM antenna reported is consisted of two divergent beams and their 3D radiation patterns are center hollowed cone. As the propagation distance increases, two beams will separate further, which requires that the location and direction of receiving antenna must be carefully modified to receive the OAM waves properly. Furthermore, for the EM waves carrying OAM with different modes radiated by the same antenna, beam divergent angle varies with the OAM mode number \(l\) [7]. In other words, at the same location, the received ...
power of signals carrying different OAM mode number will be different, which makes it more difficult to multiplex beams of different OAM modes. Thus, the above disadvantages existing in OAM antenna have severely restricted its practical application in wireless communication systems [19], [20].

In 2015, [21] proposed the concept of plane spiral orbital angular momentum waves at the first time whose 3D radiation pattern has the characteristics of side-fire, which is far more convenient for receiving and multiplexing in practical applications. This also means that compared with OAM waves transmitting along the main axis, the plane spiral OAM propagates along the transverse direction, which can be used as an effective method to solve the problem of beam divergence existing in OAM. Moreover, the PSOAM maintains the characteristics of orthogonality and spiral phase wavefront of OAM, thus, it has more considerable application prospects in the fields of PSOAM-MIMO system [22]–[24], beamforming and beam scanning [25]. So far, only four reported antennas can be capable of generating PSOAM waves. Reference [26] proposes a ring resonant cavity antenna with external horn, which has the advantage of high modal purity. Its volume is large, however, which is not conducive to the realization of dual-mode superposition. And due to its large out-of-roundness, it is still necessary to further optimize the pattern to meet the actual applications. Reference [27] proposes a ring resonant cavity antenna loaded with meta-surface, which has the disadvantages of non-miniaturized, narrow beam, and narrow impedance bandwidth. In order to solve the shortcomings of the above antennas, a dielectric resonator antenna is proposed in [28], which greatly reduces the size of the antenna, leading to a limited radiation efficiency and gain. A cylindrical conformal microstrip antenna is proposed in [25], which has sufficient reconfigurability. However, its feeding network is much larger than antenna array itself. In summary, these schemes have some defects, especially in antennas’ profile, which considerably restricts its practical applications.

Therefore, we propose a novel microstrip uniform circular array (UCA) with flat structure can generate PSOAM waves. The antenna is made up of eight microstrip Yagi antenna elements, and is fed by a self-designed feeding network. The principle of the PSOAM beam generated by the circular array antenna (CAA) is given. Then, a microstrip Yagi PSOAM CAA is designed, fabricated, and measured, which provides a new way for the design and realization of PSOAM antenna. Compared with above PSOAM antennas, this design is flat structure with the lowest profile.

II. PRINCIPLE OF GENERATING PSOAM BY MICROSTRIP YAGI CIRCULAR ANTENNA ARRAY

A. THE PRINCIPLE OF OAM GENERATED BY UNIFORM CIRCULAR ARRAY ANTENNA

The configuration of the UCA antenna is represented in Fig. 1(a). The radius of array is \( a \), and each array element is distributed in the x-y plane with the azimuth angle of \( \varphi_n = 2\pi n/N \), where \( N \) is the number of array elements, \( n = 0, 1, 2, \ldots, N - 1 \). And all array elements are fed with equal magnitude. As depicted in Fig. 1(b), the center of CAA is regarded as the phase reference point. We assume that the distance between the observation point \( P(r, \theta, \varphi) \) in far field and the phase reference point is \( r \).

The position coordinates of array element \( n \) are \( P_n = [a \cos(2\pi n/N), a \sin(2\pi n/N), 0] \), and the unit vector of signal incident direction is \( \mathbf{r} = (\sin \theta \cos \varphi, \sin \theta \sin \varphi, \cos \theta) \).

We suppose the signal arrives at the reference point earlier than the time to reach element \( n \), then the delay of array element \( n \) relative to the reference point can be expressed as:

\[
\tau_n = -\frac{1}{c} \left[ R \cos \frac{2\pi n}{N} \sin \theta \cos \varphi + R \sin \frac{2\pi n}{N} \sin \theta \sin \varphi \right] = -\frac{R}{c} \sin \theta \cos \left( \varphi - \frac{2\pi n}{N} \right) \tag{1}
\]

Thus, the corresponding phase shift can be derived as:

\[
\varphi_n(\theta, \varphi) = -\frac{2\pi}{\lambda} R \sin \theta \cos \left( \varphi - \frac{2\pi n}{N} \right), \quad n = 0, 1, \ldots, N - 1 \tag{2}
\]

We assume the excitation current is \( I_n = I e^{-j\beta_n} \), \( I \) is the current amplitude, \( \beta_n \) is the phase of the \( n \)th excitation source. According to the principle of electric field superposition in electromagnetic waves, the total field of the antenna array at the observation point is formed by the superposition fields radiated by each element. Since the array elements have the same structure, resulting in identical polarization direction in the far field, which indicates that the vector sum can be expressed as a scalar sum. Thus, the total far-field can be expressed as:

\[
E = \frac{e^{-jkr}}{r} \sum_{n=0}^{N-1} I_n [k a \sin \theta \cos(\varphi - \varphi_n) + \beta_n] \tag{3}
\]

The array factor can be obtained as:

\[
F(\theta, \varphi) = \sum_{n=0}^{N-1} I_n [k a \sin \theta \cos(\varphi - \varphi_n) + \beta_n] \tag{4}
\]
In order to obtain OAM electromagnetic wave with mode \( l \), let \( \beta_n = l\varphi_n \), then the Eq. (4) is equal to:

\[
F(\theta, \varphi) = \sum_{n=0}^{N-1} I e^{i [k l \varphi \sin \theta \cos(\varphi - \varphi_n) + l\varphi_n]}
\]

(5)

When \( N \) is sufficiently large and \( \Delta \varphi = 2\pi/N \) is small enough, the above equation can be dealt as an integral:

\[
\sum_{n=0}^{N-1} I e^{i [k l \varphi \sin \theta \cos(\varphi - \varphi_n) + l\varphi_n]}
= \frac{1}{\Delta \varphi} \sum_{n=0}^{N-1} e^{i [k l \varphi \sin \theta \cos(\varphi - \varphi_n)] \Delta \varphi}
\approx \frac{N}{2\pi} \int_0^{2\pi} e^{i [k l \varphi \sin \theta \cos(\varphi - \varphi_n)]} d\varphi_n
= Ne^{i \varphi_n} J_1(kl \varphi) = Ne^{i \varphi_n} J_1(2\pi a_1 \sin \theta)
\]

(6)

where \( J_1(2\pi a_1 \sin \theta) \) is the \( l \)th-order Bessel function of the first kind, \( a_1 = \lambda/a \), \( \lambda \) is the working wavelength of CAA in vacuum, and \( k = 2\pi/\lambda \) is wave number. It can be derived from (6) that the field radiation pattern factor contains the vortex phase factor \( e^{i \varphi} \), which demonstrates that vortex beams carrying OAM can be generated by controlling the excitation phase of the array element. The phase difference between adjacent array elements is \( 2\pi l/N \), which means that the phase of signal is delayed \( 2\pi l \) after rotating around the transmission axis once. Therefore, the radiation field also has a phase difference of \( 2\pi l \) on the \( \varphi \) plane, and then an electromagnetic wave carrying orbital angular momentum is obtained. The number of array elements has an additional influence on an OAM-generating antenna compared to a regular antenna array: it determines the largest \( l \) mode the array can generate. Namely, theory predicts \(-N/2 < l < N/2\) where \( N \) is the number of array elements [29], [3]. We have assumed that the beam axis is centered on the antenna array. In this paper, we assume \( N = 8 \). Therefore, an OAM-carrying radio beam mode numbers \( l \) = 0, ±1, ±2, ±3 can be generated if the successive phase difference from element to element is at steps 0, ±45°, ±90°, ±135°.

**B. MECHANISM OF PSOAM GENERATED BY MICRORSTRIP YAGI UCA**

Based on the above theoretical analysis, the uniform circular array antenna with a certain phase can generate vortex electromagnetic wave. Since the difference between PSOAM and OAM is the direction of beam propagation, the key to designing PSOAM antennas is to generate transverse beams while maintaining OAM characteristics. Therefore, a reasonable design is essential for transforming the OAM beams into PSOAM beams. In this paper, a special arrangement is used to focus the energy on the \( \varphi \) plane. As shown in Fig. 2, the E-plane and H-plane radiation pattern of the Yagi element are given. The half power beamwidth of E-plane is about 69°, and that of H-plane is about 119°. Because the microstrip Yagi antenna elements have an end-fire radiation pattern, and each element radiate in different azimuth angles along the radial direction. Thus, based on the mechanism of microstrip conformal antenna and horizontally polarized omnidirectional antenna array, the proposed antenna can obtain the omnidirectional radiation pattern by evenly arranging eight microstrip Yagi antennas along a circle on the xoy-plane. Fig.3 briefly demonstrates how the OAM beam is converted into PSOAM beam from the perspective of radiation pattern synthesis. Combined with the proceeding analysis on the principle of generating OAM by the UCA,
thus, the microstrip Yagi UCA antenna has potential to realize PSOAM waves by modifying the radius of antenna array reasonably.

III. ANTENNA DESIGN

A. ANTENNA GEOMETRY

Fig. 4(a) describes the total geometry of the proposed microstrip Yagi array antenna. Basically, the designed plane spiral OAM antenna contains two substrates ($S_1$ and $S_2$) and eight coaxial lines (C) between $S_1$ and $S_2$. Microstrip Yagi array elements (P) and irregular ground ($G_1$) are placed on the upper substrate $S_1$, and the feeding network (F) which consists of eight output ports is placed on the lower substrate $S_2$. The feeding network F and antenna elements (P) are connected by coaxial lines C. There are eight vias both in the substrates and the ground planes to avert contact between two ground plane ($G_1$ and $G_2$) and coaxial cores. The feeding network provides equal amplitude and constant phase difference for each adjacent array elements.

Table 1. Design parameters of the proposed array element.

| Parameter | $b_1$ | $b_2$ | $b_3$ | $c_1$ | $c_2$ | $d_1$ |
|-----------|-------|-------|-------|-------|-------|-------|
| value (mm) | 36.73 | 9.33  | 18.8  | 21.5  | 1.2   | 7     |

| Parameter | $d_2$ | $d_1$ | $d_4$ | $w_1$ | $w_2$ |
|-----------|-------|-------|-------|-------|-------|
| value (mm) | 0.8   | 4.8   | 3     | 1.8   | 3     |

The upper FR4 substrate $S_1$ is 85mm, the thickness is 1.2mm, the relative dielectric constant is 4.4, and the loss tangent is 0.02. The feeding network is arranged on the lower surface of the F4B-2 substrate $S_2$ whose radius is 40 mm, thickness is 0.8 mm, the relative dielectric constant is 2.65, and the loss tangent is 0.003. Detailed parameter values of the designed antenna element shown in Fig. 5 are listed in Table 1.

The proposed antenna is composed of eight microstrip Yagi elements, which equally distributed on the substrate plane with the radius $a$ is 35 mm. As shown in Fig. 4(b), the modal excited by eight lump ports whose phase is set up to 0°, 135°, 270°, 45°, 180°, 315°, 90°, 225°, it is used to demonstrate further with the scheme of numerical simulation by ANSYS 2019R3. The simulated radiation patterns and vortex beam for the proposed PSOAM antenna without feeding network are depicted in Fig. 6. Fig. 6 (a)-(c) show normalized E-plane (xoy), normalized H-plane (xoz) and 3D radiation pattern, respectively. It can be observed from simulated results that uniform omnidirectional radiation patterns are realized. And the cross-polarization levels are less than −30dB. The gain variations in E-plane are less than 1.6dB. Fig. 6(d) shows the vortex wave of $l = 3$ observed on a square plane. The size of observation plane is 220 mm × 220 mm, which is 200 mm away from antenna surface. We can see that the vortex phase distribution and side-fire radiation pattern are all obtained, which have verified PSOAM can be generated successfully by the proposed modal. Therefore, the simulated results of antenna are consistent with the theoretical analysis mentioned in Section II.

B. FEEDING NETWORK STRUCTURE

Uniform circular array generating vortex beam requires a battery of special feeding signals, which can satisfy identical amplitude and constant phase difference. In this paper, the feeding network of $l = 3$ is made up of one input port (0) and eight output ports (1 to 8). Fig. 7 displays its structure. On the whole, there are three types of phase-shift Wilkinson power divider in the designed feeding network. For each phase-shift power divider, it is made up of a T-shaped equal power divider, two impedance transformers ($\lambda_g$ is the guided wavelength), an isolation resistance with 100Ω and two phase shifters.

The first category of phase-shift power divider is $D_1$, which is able to realize a 540° phase difference between the two output ports with $l_{2,2} - l_{1,1} = 3/2\lambda_g$. Correspondingly, the second category of phase-shift power divider is $D_2$, which
FIGURE 6. Simulated results of the modal with lumped ports. (a) E-plane (xoy) radiation pattern. (b) H-plane (xoz) radiation pattern. (c) 3D radiation pattern. (d) vortex phase distribution with \( l = 3 \).

FIGURE 7. Structure of the feeding network for \( l = 3 \).

is \( D_3 \), which is able to realize a 135° phase difference with \( l_{3,2} - l_{3,1} = 3/8\lambda_g \).

FIGURE 8. Simulation results of feeding network. (a) Reflection coefficient, amplitude, and isolation. (b) Phase difference.

The S parameters and phase difference results between each adjacent ports of the designed feeding network for \( l = 3 \) are given in Fig. 8. The central frequency is 3.43GHz, and the bandwidth about 600MHz (3.1GHz-3.7GHz). Fig. 8(a) shows the reflection coefficient, transmission coefficient of eight output ports, and isolation with all ports. The transmission coefficients are between \(-10.08\)dB and \(-9.69\)dB with the magnitude imbalance of the equally feeding smaller than 0.39dB at 3.43GHz. Simultaneously, the isolation between port 1 and other ports are almost below 20dB, indicating a good isolation of the designed feeding network. Fig. 8(b) shows phase difference between each adjacent port is basically maintained at 135° within an error of 3.8°, which can satisfy the phase requirement of the vortex wave with \( l = 3 \).

IV. FULLWAVE SIMULATION AND MEASUREMENT

The mechanism of generating plane spiral OAM waves by a microstrip Yagi UCA antenna is interpreted in Section II. The antenna displayed in Fig. 4 was designed and simulated using ANSYS Electronics 2019R3 full-wave simulator based on finite element method, according to the parameters listed in Table 1.
As shown in Fig. 9, the archetype of the designed PSOAM antenna was fabricated. The VSWR was measured by the Agilent E5071C VNA. The E-plane radiation pattern are measured in chamber, and measurement setup in this paper is demonstrated in Fig. 10.

Fig. 11 shows the simulated and measured VSWR of the proposed PSOAM antenna. It can be seen that the simulated operation band for VSWR ≤2 are from 3.15GHz to 3.65GHz. Compared with simulated VSWR, the measured VSWR result has wider bandwidth and frequency deviation. Meanwhile, it can be seen that the measured VSWR has sudden increase at the resonance. The discrepancy is caused by the influence of the measurement environment and the tolerance of PCB processing including inaccuracy of the substrate and the welding and installing errors of the microstrip and coaxial lines. Because the distance between the upper substrate and lower substrate is close, it is difficult to weld coaxial lines, which produces additional errors. The above reasons cause other resonance modes not needed to be stimulated, so that the VSWR suddenly increases at the resonance frequency and a second resonance point appears at low frequency. Although there is a difference between the simulated VSWR and the measured VSWR, the percentage bandwidth of the latter is still wider than that of the former, which can satisfy practical requirements from 3.1GHz to 3.7GHz. Therefore, the differences are acceptable and unavoidable.

Fig. 12 represents the simulated and measured results of the far-field radiation patterns at 3.43GHz for l = 3. Fig. 12(a) shows the normalized E-plane radiation pattern whose gain variation is less than 3dB at the center frequency. Meanwhile, it can be seen from the simulated and measured results that the cross-polarization (vertical polarization) levels are about −16dB lower than the co-polarization (horizontal polarization). Compared with the results of Fig. 6, this value increases significantly, which is caused by the multi-layer structure of the Yagi antenna array. Therefore, the PSOAM antenna proposed using a multi-layer structure does increase the cross-polarization level. The simulated 3D radiation pattern whose beam propagates along the transverse direction is presented in Fig. 12(c). On the whole, the measured radiation pattern including H-plane coincide with simulated radiation pattern. The distortion of measured H-plane radiation pattern as a result of waves radiated by fabricated feeding network and inaccuracy of measurement. As depicted in Fig. 12(b), the main lobe points to the direction of θ = 90°, which demonstrates that the characteristics of a central null at θ = 0° and side-fire radiation pattern are obtained. Therefore, most of the energy is focused on side-fire direction, while the
TABLE 2. Comparison with the reported PSOAM antennas.

| Ref   | Frequency (GHz) | Impedance bandwidth | Antenna type                     | Flat size    | Profile       | Feeding method          |
|-------|-----------------|----------------------|----------------------------------|--------------|---------------|--------------------------|
| [25]  | 10              | 280MHz               | Uniform circular array           | $D_a$: 0.67 $\lambda_0$ | $0.95 \lambda_0$ | 8×8 Butler Feed network |
|       |                 |                      |                                  | $S_f$: 8.75 $\lambda_0$ * 5.72 $\lambda_0$ |               |                           |
| [26]  | 10              | <30MHz               | Traveling-Wave antenna           | $D_a$: 4.12 $\lambda_0$ | 1.66 $\lambda_0$ | 90° hybrid coupler       |
| [27]  | 15              | <30MHz               | Traveling-Wave antenna loaded with metasurface | $D_a$: 6.38 $\lambda_0$ | 0.5 $\lambda_0$ | Coaxial                  |
| [28]  | 10              | <30MHz               | Dielectric resonator antenna      | $D_a$: 0.90 $\lambda_0$ | 0.23 $\lambda_0$ | Coaxial                  |
| This paper | 3.43          | 600MHz               | Uniform circular array           | $D_a$: 1.94 $\lambda_0$ | 0.16 $\lambda_0$ | Wilkinson power divider |
|       |                 |                      |                                  | $S_f$: 0.53 $\lambda_0$ * 0.53 $\lambda_0$ |               |                           |

* $D_a$: Diameter of antenna.
* $S_f$: Size of feeding work.

Simulated peak gain is 1.9dB and the measured is about 1.7dB. It is obviously that the Yagi array antenna’s gain is much lower than that of the element. However, it is at a normal level. The reason for this phenomenon is the arrangement of the array and mutual coupling. In this design, the compact anisotropic circular array is the key factor to realize PSOAM beam. Therefore, strong mutual coupling and gain reduction are inevitable. In addition, the HPBW of the H-plane radiation pattern is 85°, which indicate that this arrangement will reduce the HPBW of the H-plane. At the same time, it can be seen that the simulated and measured radiation pattern has a few concavities compared with the simulated radiation pattern without Wilkinson power divider in Fig. 6. This difference is caused by the insertion loss of feeding network. Due to the insertion loss, the amplitude of output port is not completely consistent, while it is unavoidable and within the acceptable range.

**FIGURE 13.** Electric field phase distribution at 3.43GHz for $l = 3$.
(a) simulated result. (b) measured result.

Fig. 10 shows the test diagram of the plane near-field experiment. A Quadruple-ridged horn antenna is used as a standard near-field measurement probe, and the radiation direction of the PSOAM antenna is alignment with it. The observation plane is 1m × 1m, 1m away from the antenna. The simulated and measured results of vortex phase distribution of the proposed PSOAM antenna at 3.43GHz for $l = 3$ are shown in Fig. 13. Although some discontinuities occur in measured result and the phase singularities are deviated due to fabricating and measuring error, three spiral arms can be clearly observed in Fig. 13(b). According to the above results, we can conclude that the proposed PSOAM antenna is cable of generating the vortex beam with $l = 3$. In general, the traditional OAM beam has a phase singularity, which means that the energy null area exists in the central of OAM beam, and it will enlarge with the transmission distance increases. It will bring challenges to the reception of electromagnetic waves. Instead, for the PSOAM waves propagating along the transverse direction, its phase singularity is occupied by the antenna itself. Thus, the PSOAM antenna can ignore the problems caused by the phase singularity in practical applications.

We compare the reported PSOAM antennas in terms of center frequency, impedance bandwidth, antenna type, flat size, profile and feeding method in Table 2. We can conclude that the currently existing PSOAM antennas are limited by high profile which is the key factor restricting the integration of antenna and wireless communication system. Differently, the profile of the proposed antenna is only 0.16$\lambda_0$, which is the lowest value among the reported PSOAM antennas. Furthermore, the proposed antenna here realized PSOAM wave based on UCA, which has broad prospects for future research on reconfigurability.

**V. CONCLUSION**

In this letter, the principle of PSOAM waves generation using microstrip Yagi CAA is introduced. Then, the microstrip Yagi uniform circular array antenna generating PSOAM wave
fed by a self-designed 1-to-8 Wilkinson power divider is designed, fabricated, and measured. Different from conventional OAM antennas, eight microstrip Yagi elements are put evenly along the φ direction to focus energy on the transverse direction. Both simulated and measured results show that the proposed antenna can generate PSOAM waves. Therefore, there is no phase singularity in practical applications. Additionally, the beam divergence angle of different OAM modes will be same, which facilitates multiplexing. With the advantage of low profile, the proposed PSOAM antenna is more suitable for integration with communication system and information-rich radar system. Simultaneously, the UCA is usually applied to realize the reconfiguration of antennas, thus, the research of PSOAM antenna based on UCA will have considerable prospects in the field of reconfiguration.

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