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ABSTRACT
In this paper, we propose a reflectarray metasurface to generate two modes of orbital angular momentum (OAM) in two orthogonal directions of polarization. The metasurface comprises dual-bowtie elements of various sizes. By varying the size of the geometry, the element can simultaneously alter the phases of reflection of both polarizations. In the design, equivalent circuit models of the unit cell in both the orthogonal direction and the direction of excitation are built and discussed. These models are used to explain the results of the simulation and can help accelerate the optimization process when designing other reflective metasurfaces. Both the first- and second-order metasurface reflectarrays were simulated and measured at 10 GHz, and the results confirmed that different OAM modes can be generated in two orthogonal directions of polarization. Compared with the prevalent design, the proposed one is more compact as it has only one layer and does not require active components. We also simulated and measured the first-order metasurface reflectarray at multiple frequencies, and the results from 9.2 to 10.5 GHz show that it can generate the two polarized OAMs independently in broadband.

I. INTRODUCTION
Electromagnetic waves carry angular momentum during propagation that has two components: spin angular momentum and orbital angular momentum (OAM). The spin angular momentum relates to the states of polarization of electromagnetic waves (i.e., linear and circular polarizations), whereas the OAM corresponds to the distribution of the magnitude and wave front of the electromagnetic waves. There are a variety of ways to form electromagnetic wave beams that carry OAM in the microwave and optical frequency regions, such as spiral phase plates, metasurfaces, and patch array antennas. Given that its different modes are orthogonal to one another, the OAM has attracted considerable attention because it can be applied to communication systems to improve channel capacity without consuming extra bandwidth. In the domain of radio frequency, various methods of generating OAM beams have been proposed. Approaches to generate more than one OAM in a single device have been recently explored. In general, there are many ways to produce multiple OAM modes, including the patch antenna array, spiral antenna, resonant cavity, metasurfaces, and other planar devices. In patch antenna arrays, it is easy to design the patch antennas, but difficult to plan the feeding networks. Both the spiral antenna approach and the resonant cavity have the shortcoming of a bulky structure. The use of metasurfaces is becoming increasingly popular because of their low profile, simple feeding network, and good directivity. In 2016, Yu et al. developed three reflection metasurfaces to generate three OAM modes of vortex waves at 5.8 GHz and combined two cells aligned orthogonal to each other to achieve a dual-polarization and dual-mode OAM-generating metasurface. However, the substrate of the metasurface was multilayered and resulted in a more complex structure. In 2017, Li et al. proposed a reflectarray to independently generate reconfigurable OAM-carrying waves in orthogonal directions of polarization in radio frequency. However, the technique needs an additional source to control the active components. In 2019, Qi et al. proposed a transmissive metasurface for the generation of dual-mode polarized OAM. This structure is also multilayered with three dielectric layers and four metal surfaces. At present, techniques to obtain multiple OAM modes by designing metasurfaces are limited to the aforementioned ones, which means that the orthogonal directions of the metasurface are designed separately.
Our previous work on the single-bowtie structure took advantage of the cell’s 360° linear phase variation vs the size of the bowtie under the incidence of a wave in a polarized direction. However, the phase variation in the other orthogonal direction remains unavailable for OAM generation. In this study, we design a dual-bowtie structure to form a reflectarray metasurface to facilitate the simultaneous generation of separate OAM modes in orthogonal directions of polarization. We use x and y as the orthogonal directions. The proposed structure shows different phase responses to x- and y-polarized incident waves. Within a specific range of structure sizes, the overall phase variations are 360° and 720° in the x and y polarizations, respectively. In other words, the phase responses in both orthogonal directions of polarization are controlled by a single parameter. To the best of the authors’ knowledge, this is the first time that different phase variations for OAM generation using a single parameter have been reported. We also carried out equivalent circuit analyses to investigate the difference between the orthogonal polarizations. It is also potentially useful for accelerating the optimization process when designing other reflect metasurfaces. The results of cell structure analysis and array design were verified by simulations performed using high frequency structure simulator (HFSS) software and measurements in a microwave anechoic chamber. The proposed structure is compact, does not require active components, and has the broadband property. It can be applied further to OAM communication systems.

II. METASURFACE DESIGN

The scatter field of an array comprising M × N cells is determined by the far-field radiation patterns of the feed source and the scattering pattern of the cell. For a specified position of the unit cell of the array, the phase variations in the orthogonal directions of polarization (i.e., $\Omega_x$ and $\Omega_y$) are composed of three parts: the cell-generating phase, the OAM-related phase, and the propagating phase. The relations can be expressed as follows:

$$\left\{ \begin{array}{l} \Omega_x = \Omega_x^c + l_x \varphi + k_0 |\vec{r}_m - \vec{r}_f| \\ \Omega_y = \Omega_y^c + l_y \varphi + k_0 |\vec{r}_m - \vec{r}_f|, \end{array} \right.$$  

where $l_x$ and $l_y$ are the designed topological charges of the OAM beams, $\Omega_x^c$ and $\Omega_y^c$ are the phases generated by the cell, $\varphi$ is the azimuthal angle of the cell, $\vec{r}_m$ is the cell’s position vector, and $\vec{r}_f$ is the source position vector, as illustrated in Fig. 1.

Regarding the plane wave source, the propagating phases $(k_0 |\vec{r}_m - \vec{r}_f|)$ of the source on a metasurface are identical because the vector $\vec{r}_f$ is considered to be infinite. Consequently, the far-field phase in one direction of polarization is determined by the phase variation of a cell and its azimuthal angle on the metasurface. Given that the azimuthal angle of a cell is fixed, the phase variation in two orthogonal directions of polarization must meet the following relation to generate two modes in the x and y directions:

$$\Omega_x^c / \Omega_y^c = l_x / l_y.$$  

Consider $l_x = 1$ and $l_y = 2$ as an example. The phase variation in the y polarization direction is expected to be exactly twice that in the x polarization direction only when the condition $\Omega_y^c = 2 \Omega_x^c$ is met. This condition serves as a guideline for cell design.

III. DESIGN OF DUAL-BOWTIE UNIT CELL

The proposed unit cell consists of two bowtie-shaped structures placed orthogonally and surrounded by two rings. As shown in Fig. 2, the unit cell of the reflective metasurface is a three-layered structure. The top layer is the proposed dual-bowtie structure, the middle layer is a dielectric substrate, and the bottom layer is pure metal that acts as a perfect electric conductor. We have previously proposed a single bowtie ring structure with linear phase response. An additional 360° of phase variation can be achieved with the addition of an extra ring surrounding the single-ring structure. Therefore, two bowtie structures are placed orthogonal to
FIG. 3. Simulation setup (a) for simulation and resultant simulated phase variation (b) in the x and y polarizations. $M_1$ and $M_2$ are master boundary conditions, while $S_1$ and $S_2$ are the corresponding slave boundary conditions. Modes 1 and 2 of the Floquet port are the x and y polarizations, respectively. In the phase and magnitude of the reflect coefficient, the solid and hollow scatter lines represent the x- and y-polarization fields, respectively. $P_x^1$ and $P_x^2$ are the resonance peaks of the x polarization, while $P_y^1$ and $P_y^2$ are those of the y polarization.

We used the master and slave boundary conditions and a Floquet port at 10 GHz for the HFSS simulations. The Floquet port has two orthogonal modes, x and y polarizations, as presented in Fig. 3(a).

The proposed unit cell has phase ranges of $360^\circ$ and $720^\circ$ in the x and y directions of polarization, respectively, whereas the geometric parameter $r_1$ of the dual-bowtie structure varies from 2.125 mm to 3 mm, as presented in Fig. 3(b). Note that although the cell size varies by less than 1 mm, the manufacture tolerance of the PCB (printed circuit board) is 0.05 mm while the minimal width of the two rings is 0.84 mm. Given that the manufacture tolerance is significantly smaller than the minimal width, it is sufficient for the manufacture process to ensure accuracy. The two phase response curves were nearly linear. The resultant satisfactory performance in terms of magnitude is also shown in Fig. 3(b), that is, the magnitudes of both polarizations were larger than $-1$ dB, thus guaranteeing a uniform distribution of the magnitude of the generated OAM beams. As shown in Fig. 3(b), the resonance peaks were at 2.08 mm and 2.62 mm in the x polarization direction and at 2.46 mm and 2.77 mm in the y polarization direction.

IV. EQUIVALENT CIRCUIT MODEL ANALYSIS

To understand the working principle of the proposed dual-bowtie unit cell, we carried out an in-depth analysis of the structure to extract the corresponding circuit parameters. The dual-bowtie structure is composed of two bowtie shapes arranged orthogonally and two outer rings, where the outermost ring is carved to form a gap. Given that the two rings are curved, they cannot be accurately modeled as capacitance and inductance. We divided each into eight equal sectors, with each sector approximated by a rectangular patch.
linear with the geometric parameter selected here. For the inductor, the length of the metal patch was \( l \), its width was \( w \), its thickness was \( t \), the distance between the metal patches was \( w \), and the geometric equivalent distance (GMD) was defined as the distance between the geometric centers of parallel metal patches. The variable GMD of self-inductance can be expressed as

\[
I = \frac{\mu_0 l w}{2 \pi \text{GMD}}
\]

TABLE I. Values of capacitance (unit: pF/mm) and inductance (unit: nH/mm) in the equivalent circuits.

| Name | \( C_{x1} \) | \( C_{x2} \) | \( C_{x3} \) | \( L_{x1} \) | \( L_{x2} \) | \( L_{x3} \) | \( L_{x4} \) |
|------|-------------|-------------|-------------|-------------|-------------|-------------|-------------|
| Value | 0.1288      | 0.0220      | 0.1836      | 0.1020      | 0.5940      | 0.0688      | 0.4016      |

| Name | \( L_{x5} \) | \( L_{x6} \) | \( C_{y1} \) | \( C_{y2} \) | \( C_{y3} \) | \( C_{y4} \) | \( L_{y1} \) |
|------|-------------|-------------|-------------|-------------|-------------|-------------|-------------|
| Value | 0.4524      | 0.8120      | 0.1288      | 0.0220      | 0.1836      | 0.0316      | 0.1020      |

| Name | \( L_{y2} \) | \( L_{y3} \) | \( L_{y4} \) | \( L_{y5} \) | \( L_{y6} \) |
|------|-------------|-------------|-------------|-------------|-------------|
| Value | 0.5940      | 0.0688      | 0.3108      | 0.7084      | 0.4016      |

FIG. 5. Equivalent circuit models in the x (a) and y (b) directions of polarizations. An additional capacitance \( C_{y4} \) is introduced in the y direction of polarization in (b). x (c) and y (d) polarization of the magnitude of the partial component of the E field that is tangential to the XOY plane, i.e., the surface magnetic current density (units: V/m).

FIG. 6. The phase and amplitude obtained in both directions of polarization of the equivalent circuit models.
\[ \ln GMD_{\text{self}} = \ln \left( \sqrt{w^2 + t^2} \right) - \left( \frac{1}{6} \right) \left( \frac{w^2}{t^2} \right) \ln \sqrt{1 + \left( \frac{t^2}{w^2} \right)} \\
+ \left( \frac{t^2}{w^2} \right) \ln \sqrt{1 + \left( \frac{w^2}{t^2} \right)} + \left( \frac{2}{3} \right) \left( \frac{w}{t} \right) \tan^{-1} \left( \frac{t}{w} \right) \\
+ \left( \frac{t}{w} \right) \tan^{-1} \left( \frac{w}{t} \right) - \left( \frac{25}{12} \right), \quad (3) \]

and self-inductance can be calculated as follows:

\[ L = \left( \mu_0 / 2\pi \right) \ln \left[ \left( 1 / GMD \right) + \sqrt{1 + \left( GMD^2 / F^2 \right)} \right] \\
- \sqrt{1 + \left( GMD^2 / F^2 \right)} + \left( GMD / l \right). \quad (4) \]

For the capacitor, the length of the capacitor plates was \( l \), their thickness was \( w \), the spacing between the plates was \( s \), and the thickness of the medium was \( h \). We considered only the odd mode, which can be expressed as

\[ C_0 = C_p + C_f + C_{ga} + C_{gd}, \quad (5) \]

where \( C_p = \varepsilon_0 \varepsilon_r w / h \), \( C_f = \left( \sqrt{\varepsilon_r / \varepsilon_0} - C_p \right) / 2 \), \( C_{ga} = \varepsilon_0 K(k') / K(k) \), \( k = (s/h) / (s/h + 2w/h) \), and \( k' = \sqrt{1 - k^2} \).

\[ \frac{K(k)}{K(k')} = \begin{cases} (1/\pi) \ln \left[ 2 \left( 1 + \sqrt{k'} \right) / \left( 1 - \sqrt{k'} \right) \right], & 0 \leq k^2 \leq 0.5 \\
\pi / \left[ \ln \left[ 2 \left( 1 + \sqrt{k} \right) / \left( 1 - \sqrt{k} \right) \right] \right], & 0.5 \leq k^2 \leq 1, \end{cases} \quad (6) \]

and \( C_{gd} \) can be written as follows:

\[ C_{gd} = \left( \varepsilon_0 \varepsilon_r / \pi \right) \ln \left[ \coth \left( \pi s / 4h \right) \right] + 0.65 C_f \left[ 0.02 h \sqrt{\varepsilon_r / s} + 1 - \varepsilon_r^{-2} \right]. \quad (7) \]

Because the shape of the bowtie was irregular, it was treated as equivalent to a rectangle for convenience of calculation, as shown in Fig. 4. The bowtie in the orthogonal direction could be considered a capacitance-inductance-capacitance (CLC) series branch on the central bowtie. Because the minimum distance between the orthogonal bowties was small, the capacitance was too large to consider here.
Accordingly, the CLC series branch was considered as two inductors in parallel with the inductance of the bowtie structure, as shown in Fig. 5.

The equivalent circuit diagrams in the x and y directions of polarization are presented in Fig. 5. $L_{x(1)}$ and $L_{x(2)}$ are the equivalent inductors of the outer ring, and $L_{x(3)}$ and $L_{x(4)}$ represent the inductors of the inner ring. $L_{y(1)}$ stands for the half-bowtie in the orthogonal direction of the polarization, while $L_{y(2)}$ symbolizes the central bowtie. $C_{x(1)}$ and $C_{x(2)}$ are the capacitances between the rings, and the capacitance between the inner ring and the central bowtie is modeled as $C_{y(3)}$. The most significant difference between the polarizations in the x and y directions is one more capacitor $C_{y4}$ in the outer ring in the polarization in the y direction. It results in the noticeable difference that the E-field between the inner ring and the outer ring in Fig. 5(d) is not as symmetric as that in Fig. 5(c).

Other differences persisted mainly in the equivalent inductances and capacitances due to different geometric parameters. We chose variable $r_1$ in Fig. 2 as the indicator of cell size. The capacitance and inductance can then be expressed as

$$C = C_n * r_1, \quad L = L_n * r_1,$$

and the calculated $C_n$ and $L_n$ are listed in Table I.

The equivalent circuit was simulated using Advanced Design System (ADS) software, and the phase and amplitude obtained in both directions of polarization are shown in Fig. 6. The result of the phase was consistent with the phase characteristics simulated in HFSS. There were two periodic phase changes in each of the x and y directions of polarization, but in the range from 1.25 mm to 3.75 mm, the x-phase changed only one cycle, whereas the y-phase had two cycles. Note that the positions of resonance in the results of the simulated circuit were not identical to those of the simulation of the cell in HFSS because this was only a coarse evaluation, and some approximations were made to simplify the analysis of the circuit model.

It is useful to know the parts of the circuits that affect the two resonance positions in both directions of polarization. In the x polarization direction, we chose $L_{ring}$ as the inductor of the approximate rectangle, and

$$L_{x3} = L_{ring} * \cos^2(3\pi/8), \quad L_{x4} = L_{ring} * \cos^2(\pi/8),$$

and chose $L_{bowtie}$ as the inductor of the bowtie. Although both $L_{ring}$ and $L_{bowtie}$ are determined by the sizes of the equivalent rectangles shown in Fig. 4, we changed their values in ADS to investigate their effects on the resonances of the structure.

Figure 7 shows the different S11 curves simulated by the ADS software when $L_{ring}$ and $L_{bowtie}$ were changed. The curves in Fig. 7(a) show that both the resonance peaks shifted to the left when $L_{ring}$ increased. We also simulated the current distributions at different
resonance points for orthogonal directions of excitation, and the results are shown in Fig. 7 as subgraphs. The current distribution of the first resonance peak for the x polarization obtained using HFSS software is shown in the subgraph of Fig. 7(a), from which it is clear that the largest current was located in the inner ring. However, the outer ring and central bowtie also revealed noticeable current distributions. The simulated S11 curves in Fig. 7(a) confirm that both resonance peaks were affected by $L_{\text{ring}}$. This is easy to understand because a resonant circuit is formed by the two rings, and another circuit is constructed by the rings and the bowtie. By contrast, the HFSS-simulated current distribution for the second resonance peak in the subgraph of Fig. 7(b) shows that the maximum current was in the central bowtie. When we changed the value of $L_{\text{bowtie}}$, the second resonance peak shifted while the first resonance peak stayed in the same position. Analogous observations apply to the case of y polarization, the results of which are shown in Figs. 7(c) and 7(d).

Considering the circuit along the direction of the y polarization, the most significant difference between this circuit and the equivalent circuit was the capacitance $C_y$ introduced by the gap of the outer ring. The result of a simulation of the circuit for the y polarization while neglecting $C_y$ is shown in Fig. 8, from which it is clear that without the gap capacitance, the curve of the magnitude was coincident with the result of the case of x polarization, except at the second resonance peak. This is because the bowtie for the y polarization was wider and $L_{\text{bowtie}}$ was smaller than that in the x polarization circuit. When $C_y$ was considered, both resonance peaks shifted toward the larger cell size, and the gap capacitance affected the first resonance peak to a greater extent than the second. Eventually, different phase responses were formed.

Although this equivalent circuit model is simple and coarse, it can capture the main features of the metasurface cell and can very quickly simulate them using the ADS software. This is useful in designing a metasurface because the full-wave simulation software HFSS takes significantly longer to simulate even one realization of the metasurface. Thus, to save time in the optimization process, we can use this equivalent circuit model to analyze modes of variation, e.g., to determine in real time the trends of variation of the parameter S when changing some geometric parameters (e.g., $g_1$ in Fig. 2 that affects $L_{\text{ring}}$ and $\theta_1$, $\theta_2$, $r_1$, and $r_3$ that affect $L_{\text{bowtie}}$) as in Figs. 7 and 8.
FIG. 12. Experimental setup of the dual OAM measurement. The probe was positioned 900 mm from the metasurface. The scanning plane was 500 mm × 500 mm due to limitations of the scanning equipment.

V. RESULTS

Based on the analysis of the dual-bowtie structure, we designed two reflective metasurfaces composed of dual-bowtie unit cells aligned in the same direction. The azimuthal angle of the cell’s position and the phase required to generate the OAM vortex waves were considered to initially obtain the geometric size of the cell and calculate the required phase response of cells. We obtained the size of the cell using the phase response as shown by the curves in Fig. 3(b). The distribution of $r_1$ corresponding to the cell is shown in Fig. 9(a) (one-order array) and Fig. 9(b) (two-order array).

The metasurface was a 17 × 17 array with the center cell and cells in the corners removed, thus forming a circular surface with a radius of 135 mm. Taconic RF-45 was used as dielectric substrate. It was 3.2 mm thick and had a dielectric constant of 4.5. On the basis of theoretical analysis, the metasurface was determined to be able to generate different OAM modes when the incident waves were in different states of polarization.

The HFSS software was used to verify the reflectarray design. The zero-phase of the plane wave feed was set to $1\lambda$ above the metasurface, while the wave vector of the feed was set to vertically point toward the metasurface. The observation plane was located at $60\lambda$ above the metasurface with dimensions of $24\lambda$ × $24\lambda$. The results of the numerical simulation of the magnitude and phase obtained from the observation plane are presented in Fig. 10.

Figure 10 shows that the OAM modes of $l = 1$ for the x direction of polarization and $l = 2$ for the y direction of polarization were successfully generated with the first-order metasurface, whereas the OAM modes of $l = 2$ for the x polarization and $l = 4$ for the y polarization were produced with the second-order metasurface. Regarding the magnitude distribution in Fig. 3(b), both OAM modes shared the characteristic of having a hole in the center, which is a typical pattern of magnitude for OAM vortex waves. Notably, the size of the hole was larger when the number of OAM modes was larger.

Samples of the reflective array metasurfaces were manufactured to investigate the accuracy of our proposed reflective array

FIG. 13. Measured magnitude [(a)–(d)] and phase [(e)–(h)] of reflection fields of the x-polarization [(a), (c), (e), and (g)] and y-polarization [(b), (d), (f), and (h)]. [(a) and (c)] Magnitude distribution of the x-polarization field with OAM modes 1 and 2, [(b) and (d)] magnitude distribution of the y-polarization field with OAM modes 2 and 4, [(e) and (g)] phase distribution of the x-polarization field with OAM modes 1 and 2, and [(f) and (h)] phase distribution of the y-polarization field with OAM modes 2 and 4.
design, as shown in Fig. 11. Further experiments were conducted by using a microwave anechoic chamber. The measurement setup in the microwave anechoic chamber is shown in Fig. 12.

A low-side lobe lens horn antenna with a working frequency range of 8.2–12.4 GHz was used as plane wave generator. A probe was placed on a measurement plane 900 mm (i.e., 30 wavelengths at 10 GHz) from the reflectarray metasurface to measure

FIG. 14. Simulation results of x-polarization and y-polarization at different frequencies.

FIG. 15. Experimental results of x-polarization and y-polarization at different frequencies.
Owing to the nearly linear characteristics of the phase variation curves in Fig. 3(b), our design exhibited the broadband property. We use the first-order array as an example. Its phase distribution in the simulation is shown in Fig. 14 and experimental result is shown in Fig. 15. Within a frequency range of 9.2 GHz–10.5 GHz, the characteristics of the OAM phase were well maintained and the broadband property confirmed the satisfactory performance of our design.

VI. CONCLUSION

A method to generate two modes of the OAM in two orthogonal directions of polarization was introduced in this paper by considering that the dual-bowtie structure exhibits different phase responses in orthogonal directions of polarization while maintaining linear phase characteristics. A theoretical analysis of the phase variation vs. cell size verified the formation of the metasurface array, which allowed the reflective array to control two OAM modes with one degree of freedom. Equivalent circuit models for both orthogonal directions of polarization were discussed and used to explain the results of simulations and can be used to further optimize the cell structure. They can also be used for accelerating the optimization process when designing other reflect metasurfaces. The samples were fabricated, and an experiment was conducted using a microwave anechoic chamber. A comparison of the results of the simulations and experiment verified the accuracy of the designed metasurface. The broadband property further confirmed the satisfactory performance of our design. On the whole, the dual-bowtie cell can control two OAM modes in orthogonal directions of polarization with only one degree of freedom. The proposed structure and design procedure can be used as reference for an OAM multiplexing communication device as well as other aspects of OAM communication systems.

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REFERENCES

1. T. Watanabe, M. Fujiy, T. Watanabe, N. Toyama, and Y. Iketa, Rev. Sci. Instrum. 75(12), 5131–5135 (2004).
2. N. Yu, P. Genevet, M. A. Kats, F. Aieta, J.-P. Tetienne, F. Capasso, and Z. Gaburro, Science 334(6054), 333–337 (2011).
3. Q. Liu, Z. N. Chen, Y. Liu, F. Li, Y. Chen, and Z. Mo, IEEE Trans. Antennas Propag. 66(4), 1794–1804 (2018).
4. Y. Yang, C. Zhang, H. F. Ma, J. Zhao, Y. J. Dai, W. Yuan, L. X. Yang, Q. Cheng, and T. J. Cui, Appl. Phys. Lett. 112(20), 203501 (2018).
5. S. Thidé, H. Then, J. Sjöholm, K. Palmer, J. Bergman, T. D. Carozzi, Y. N. Istromin, N. H. lVarginov, and R. Khantimov, Phys. Rev. Lett. 99(8), 087701 (2007).
6. F. Tamburini, E. Mari, A. Spenseli, B. Thidé, A. Bianchini, and F. Romanato, New J. Phys. 14(3), 33001 (2012).
7. R. Niemiec, C. Brousseau, K. Mahdjoubi, O. Emile, and A. Menard, IEEE Antennas Wireless Propag. Lett. 13, 1011–1014 (2014).
8. Q. Bai, A. Tennant, and B. Allen, Electron. Lett. 50(20), 1414–1415 (2014).
9. F. Tamburini, C. Barbieri, F. Romanato, E. Mari, and B. Thidé, Appl. Phys. Lett. 99(20), 204102 (2011).
10. L. Cheng, W. Hong, and Z.-C. Hao, Sci. Rep. 4(1), 4814 (2014).
11. A. Tennant and B. Allen, Electron. Lett. 48(21), 1365–1366 (2012).
12. S. M. Mohammadi, L. K. S. Daldorff, J. E. S. Bergman, R. L. Karlsson, B. Thide, K. Forooshesh, T. D. Carozzi, and B. Isan, IEEE Trans. Antennas Propag. 58(2), 565–572 (2010).
13. R. Swain and R. K. Mishra, AIP Adv. 9(1), 015101 (2019).
14. Y. Z. Ran, J. G. Liang, T. Cai, W. Y. Ji, and G. M. Wang, AIP Adv. 8(9), 095201 (2018).
15. D. D. Liu, L. Q. Gui, C. Zhou, Z. X. Zhang, H. Chen, and T. Jiang, AIP Adv. 7(9), 095113 (2017).
16. B. Cheng, L. Wu, and S. H. Tao, AIP Adv. 6(8), 085322 (2016).
17. B. Liu, Y. Cui, and R. Li, IEEE Antennas Wireless Propag. Lett. 16, 744–747 (2017).
18. Y. Huang, Z. Qi, Q. Li, H. Zhu, X. Li, X. Li, and X. Jiang, in 2018 IEEE Asia-Pacific Conference on Antennas and Propagation (APCAP) (IEEE, 2018), pp. 494–495.
19. Y. M. Zhang and J. L. Li, IEEE Antennas Wireless Propag. Lett. 17(4), 719–721 (2018).
20. X.-D. Bai, X.-L. Liang, Y.-T. Sun, P.-C. Hu, Y. Yao, K. Wang, J.-P. Geng, and R.-H. Jin, Sci. Rep. 7(1), 40099 (2017).
21. F. Tamburini, E. Mari, A. Sponselli, B. Thide, A. Bianchini, and F. Romanato, in 2018 IEEE Asia-Pacific Conference on Computing, Mathematics and Engineering Technologies (CoMET) (IEEE, 2019), pp. 1–4.
22. Z. A. Asel, S. Guili, S. Wenhui, and D. Fan, in 2019 2nd International Conference on Computing, Mathematics and Engineering Technologies (CoMET) (IEEE, 2019), pp. 1–4.
23. Z. Zhang and S. Xiao, in 2016 IEEE MTT-S International Wireless Symposium (IWS) (IEEE, 2016), pp. 1–4.
24. H. Li, L. Kang, F. Wei, Y.-M. Cai, and Y.-Z. Yin, IEEE Antennas Wireless Propag. Lett. 16, 3022–3025 (2017).
25. K. Yuri, N. Homma, and K. Murata, IEEE Trans. Antennas Propag. 67(7), 4878–4882 (2019).
26. R. Gaffoglio, A. Caglierio, A. D. Vita, and B. Sacco, Radio Sci. 51(6), 645–658, https://doi.org/10.1002/2015rs003862 (2016).
27. Y.-G. Guo and G.-M. Yang, IEEE Antennas Wireless Propag. Lett. 16, 404–407 (2017).
28. D. Wei, Z.-G. Guo, and Y.-G. Yang, in 2018 8th International Conference on Signal Processing and Communication Systems (ICSPCS) (IEEE, 2018), pp. 1–4.
29. X. Hui, S. Zheng, Y. Chen, Y. Hu, X. Jin, H. Chi, and X. Zhang, Sci. Rep. 5(1), 10148 (2015).
30. A. Sawant and E. M. Choi, Phys. Plasmas 24(11), 113109 (2017).
31. G. Junkin, IEEE Trans. Antennas Propag. 67(3), 1459–1466 (2019).
32. Y. Zhang, Y. Lyu, H. Wang, X. Zhang, and X. Jin, IEEE Antennas Wireless Propag. Lett. 17(1), 172–175 (2018).
33. Y. Zhang, H. Wang, D. Liao, and W. Fu, Sci. Rep. 8(1), 2970 (2018).
34. F. Qin, S. Gao, W.-C. Cheng, Y. Liu, H.-L. Zhang, and G. Wei, IEEE Access 6, 61006–61013 (2018).
35. X. Meng, J. J. Wu, Z. S. Wu, T. Qu, and L. Yang, J. Phys. D: Appl. Phys. 52(30), 305002 (2019).
36. X. Qi, Z. Zhang, X. Zong, X. Que, Z. Nie, and J. Hu, Sci. Rep. 9(1), 97 (2019).
37 C.-C. Li, L.-S. Wu, and W.-Y. Yin, AIP Adv. 8(5), 055331 (2018).
38 S. Yu, L. Li, and G. Shi, Appl. Phys. Express 9(8), 82202 (2016).
39 S. Yu, L. Li, G. Shi, C. Zhu, X. Zhou, and Y. Shi, Appl. Phys. Lett. 108(12), 121903 (2016).
40 S. X. Yu, L. Li, G. M. Shi, C. Zhu, and Y. Shi, Appl. Phys. Lett. 108(24), 241901 (2016).
41 N. Kou, S. X. Yu, and L. Li, Appl. Phys. Express 10(1), 016701 (2017).
42 Y. Xu, Z. Guo, and G. Yang, Microw. Opt. Technol. Lett. 61(10), 2392–2398 (2019).
43 P. Nayeri, F. Yang, and A. Z. Elsherbeni, Reflectarray Antennas: Theory, Designs, and Applications (John Wiley & Sons Ltd., 2018).
44 H.-M. Hsu, IEEE Trans. Electron Devices 51(8), 1343–1346 (2004).
45 R. Garg and I. J. Bahl, IEEE Trans. Microwave Theory Tech. 27(7), 700–705 (1979).