Since orbital angular momentum (OAM) being investigated intensively in the optical region, there have been growing interests in employing OAM to solve the problem in wireless communications as a new method. It is found that the independence between different OAM modes is crucial to wireless communications. Motivated by the tremendous potential of OAM in communication systems, a novel method to generate vortex beams by spoof surface plasmon polariton (SPP) is proposed. A looped double-layer spoof SPP waveguide is applied to realize the transmission of electromagnetic waves. Beam emitting is accomplished through a series of circular patches, whose role is not only the radiation units but also resonators giving rise to the phase shifts required by the vortex beam. The proposed method is validated by both numerical simulation and experiment. The measured results show that spoof SPP are radiated by the circular patches and vortex beams carrying different OAM modes are observed at different frequencies. The proposed method possesses smaller size and is much easier to be integrated into systems and integrated circuits. The simple structure and design procedure make the proposed method promising in future wireless communication systems.

1. Introduction

Optical vortices are light beams with helical phase fronts and azimuthal components of wave vectors. Since the fact that photons in optical vortices carry orbital angular momentum (OAM) was discovered,[1] OAM modes have been investigated intensively in the field of optical microscopy,[2] micromanipulation,[3–7] super-resolution imaging,[8,9] and quantum information technologies.[10–15] After a series of profound studies, there have been growing interests in employing OAM to solve the problem in wireless communications as a new method. The first radio OAM mode (or twisted wave) simulation was performed in 2007, providing a theoretical foundation for OAM-based wireless communications.[16] In 2012, the first experiment of wireless radio transmission applying the vortex beam was completed, which verifies that the OAM-carried electromagnetic (EM) waves make contributions to the increase of communication capacity without increasing the bandwidth.[17] It has also been demonstrated that the EM waves with different OAM modes can work independently to each other in the wireless communications. That is to say, if the mode of the transmitting OAM wave is designed as l, spiral phase plates that generate -l mode OAM should be applied to receive the Gaussian beam by the receiving antenna.[18]

Owing to novel advantages of OAM modes, the approaches to generate EM waves carrying OAMs become a focused research area. Up to now, there have been many methods to generate OAMs. The most attractive way among them is to use spiral phase plate due to its simple structure and practicality,[5,19–22] which has been widely adopted in optics, followed by the microwave and millimeter-wave bands. The spiral phase plate is such a diffractive element that can modulate the phase of EM fields by its optical thickness. The other widely used method is antenna array.[23–27] Although the theory and technology foundation have already almost been perfect, the complex phase-shifting network is necessary to generate the required rotation phase. Not only the phase relation among the radiation units should be guaranteed, but also the power needs to be consistent to reach high-quality OAMs. When the mode number of OAM is larger, more antennas are needed, which will result in significant increase of complexity and cost.
Recent progresses in metasurfaces offer new ways of manipulating phase distributions of EM waves. By properly arranging subwavelength unit cells with different dimensions on metasurface, the EM waves interacted with the metasurface can be translated into the one with the OAM mode, and the total phase range of the metasurface determines the mode number of OAMs. It is noteworthy that some significant contributions have been reported as the basis of metasurface for OAM generations and detections. In 2011, the research on light propagations with phase discontinuities makes it possible to reach full-phase (360°, or 2π) coverage, providing significant ability to manipulate the EM waves. Then a series of pioneering methods to generate and detect OAMs based on phase discontinuities has been proposed. Meanwhile, more and more attention has been paid on single cavities due to the simple structure and easy fabrication, such as whispering-gallery mode resonators. The structures supporting whispering-gallery modes have interactions with the radiation mode, leading to the appearance of OAM modes. Half-mode substrate integrated waveguide antenna was also demonstrated to generate vortex beams. There are advantages of these existing methods but also disadvantages, such as the complexity of structures, the difficulty in controlling OAM modes, and the limited number of OAM modes.

One issue that should be noted for the available methods is the system integration. For example, the spiral plate is often used as an additional component covering the emitter for generating OAMs, making the system have big volume. This issue can be neglected for optical systems, but not for low frequency systems. In recent years, spoof surface plasmon polaritons (SPP) have been investigated to produce compact microwave components to enhance the performance of information systems. The proposed SPP-based vortex beam emitter in this manuscript is one of these attempts. The unique characteristic of OAM can bring large communication capacity for the communication systems. Most importantly, SPP-based vortex beam emitter can be easily integrated to the communication systems. Due to this consideration, we study the vortex beam emitter using planar spoof SPP.

In this article, a novel method is proposed to generate the vortex beams based on spoof SPP. A looped double-layer spoof SPP waveguide is applied to construct the transmission route of the EM waves, while a series of circular patches is set beside the spoof SPP waveguide for beam emitting. At the same time, the circular patches also function as resonators modulating the phase of radiation beam. The theoretical calculation of the beam emission is illustrated, while the prototype is simulated and measured. The measured results agree very well with the simulations, which are also predicted by the theoretical analysis. We show that the vortex beams with different OAM modes are generated at different frequencies without any changes in structure. High purity of the expected mode can be observed clearly at every resonant frequencies. Moreover, the proposed structure can be easily integrated into systems and integrated circuits (ICs) as an important part of the wireless communications. To the best of our knowledge, this is the first time to produce OAM modes using spoof SPP. The proposed method not only expands the application area of spoof SPP, but also provides a much easier way to generate vortex beams.

Figure 1. a) Prototype of the vortex beam emitter, in which the yellow part is modeled as copper in the simulation and the blue part is the dielectric substrate. Particularly, the lighter yellow denotes the structure on the top of substrate, while the darker yellow represents the structure on the bottom of substrate. b) Detailed illustration of the connection between the top and bottom structures.

2. Results

2.1. Structure and Operating Principle

Since the vortex beam owns helical phase fronts and azimuthal component of wave vector, the key to generate the vortex beam is the rotated phase distribution and radiation. Figure 1a illustrates a prototype of the proposed structure, in which a single-side corrugated metallic strip is used to form the spoof SPP waveguide. Because of the loop structure, the spoof SPP waveguide is designed on both sides of the dielectric substrate in order to avoid the overlap. The top and bottom SPP waveguides are connected by a metallic via, as shown in Figure 1b. It was declared that the metallic via can form surface mode on the transition part, instead of the waveguide mode caused by boundaries of the two layers.

This results in the high transmission efficiency, especially when the thickness of the substrate is large. The substrate in this particular design is chosen as commercial printed circuit board, F4B, with a relative permittivity of 2.65 and loss tangent of 0.003. To reduce the effect coming from the overlap between top and bottom SPP waveguides, the thickness of the substrate is set as 3 mm. The spoof SPP waveguide is made to be a loop with radius of 80 mm, for the purpose to form the whole length as an integer times of wavelength at the expected radiation central frequency (6 GHz). Here, a series of circular metal patches is used as radiators. When the circular patches are placed around the spoof SPP waveguide, it has been demonstrated that the SPP waves are easily coupled to the patches for efficient radiations.

The radius of the circular patch is set as 8 mm for the same sake, so that the single-pass phase shift after a circular resonator can achieve 2π, which will be explained in details later.

To understand the working principle of the proposed OAM emitter, we firstly consider a spoof SPP waveguide composed of the unit cell shown in Figure 2. The detailed size of the spoof SPP is designed as p = 5 mm, w = 5 mm, a = 2 mm, and h = 4 mm. The dispersion curve is obtained through numerical method. From the dispersion curve, it can be seen obviously that
the spoof SPP waveguide owns different propagating constants ($k$) at different frequencies. That is to say, when the EM waves propagate along the spoof SPP waveguide for the same physical distance, the phase shift varies at different frequencies. If the spoof SPP waveguide is made to be a loop, the rotated phase distribution will be achieved along the spoof SPP waveguide and a phase gradient along the azimuth direction will be obtained. According to the principle of OAM mode index, the mode index of the traveling-wave loop structure can be calculated as

$$l' = \frac{2\pi R}{\lambda g} = \frac{MR\pi}{p},$$

where $R$ is the radius of the loop structure, $p$ is the period of the spoof SPP unit cell, and $M$ is defined as $M = \frac{k_0 p}{\pi}$, which can be read form Figure 2. Taking the central frequency considered in this design (6 GHz) as example, we obtain the mode index corresponding the structure without circular patches as $l' = 15$. This mode index of the spoof SPP waveguide without patches is quite different to the obtained OAM mode index, and will be explained later.

When the circular patches are put beside the spoof SPP waveguide, a phase lag appears along with the energy coupling. To illustrate this issue, we consider a single circular metal patch placed beside the SPP waveguide, as shown in Figure 3a. The relationship among the input electric field $E_1$, the output electric field $E_2$, and the circulatory electric fields $E_3$ and $E_4$ can be established according to the coupling relation and the electric field in the feedback route:

$$\begin{pmatrix} E_2 \\ E_4 \end{pmatrix} = \begin{pmatrix} \rho & i\kappa \\ -i\kappa & \rho \end{pmatrix} \begin{pmatrix} E_1 \\ E_3 \end{pmatrix}$$

(1)

where $\rho$ and $\kappa$ are the self-coupling and mutual-coupling coefficients, respectively. Both factors are supposed to be independent of frequency and satisfy $\rho^2 + \kappa^2 = 1$. The circulatory electric field $E_3$ becomes $E_4$ after the propagation through the feedback route, as depicted as:

$$E_3 = e^{-i\phi} E_4 \Rightarrow E_4 = a e^{i\phi} E_3$$

(2)

where $\alpha$ is the attenuation of the circular resonator, $\phi$ denotes the single-pass phase shift, $a$ is the single-pass amplitude transfer factor, $\tau$ signifies the single-pass transition time, and $\omega$ is the frequency. From Equations (1) and (2), the relationship between the input electric field $E_1$ and output electric field $E_2$ can be obtained as:

$$t = \frac{E_2}{E_1} = \frac{\rho - ae^{i\phi}}{1 - \rho ae^{i\phi}}$$

(3)

Then the phase lag introduced by the circular resonator is determined by $\Phi = \arg (t)$.

In order to get an intuitive recognition of the phase lag introduced by the circular patches, we suppose that the single-pass amplitude transfer factor is 1. Then the effective phase lags after the circular resonator under different self-coupling coefficients are presented in Figure 3b. We note that, no matter what the self-coupling coefficient is, the effective phase-shift difference between the maximum single-pass phase shift and the minimum single-pass phase shift keeps $2\pi$ all the time. That is to say, if the single-pass phase shift after the circular resonator is made to be $2\pi$, then the effective phase-shift difference between the situations with and without the circular resonator achieves $2\pi$. This conclusion is necessary to design the OAM structures. To verify our theory, the phase changes along the spoof SPP waveguides without and with the circular patches are simulated. We draw a line along the two kinds of spoof SPP waveguides, as shown by the red lines in Figures 4a and 4c, and then monitor the phase along the lines. The simulated results are given in Figures 4b and 4d. We clearly observe from Figure 4b that the phase exhibits regular performance without the circular patch. While the patch is added to the spoof SPP waveguide, there is a sudden reverse change at the location where the patch appears, and the
phase change is about $2\pi$, as illustrated in Figure 4d. That is to say, when there is a circular resonator, there is a $2\pi$ phase lag, verifying the condition mentioned before.

Therefore, the final radiation phase is a combination of the phase from the spoof SPP waveguide and the phase lag brought from the circular metal patches. As a consequence, if the total phase shift around the loop is $l$ times of $2\pi$, then the OAM mode number is $l$.

In this particular design, there are 15 patches in total, resulting in a total phase lag of 15 times of $2\pi$. Combined with the phase provided by the spoof SPP waveguide, we conclude that the mode number ($l$) of the radiation beam around 6 GHz is zero. If the OAM mode number ($l$) is expected as 1, the phase provided by the spoof SPP waveguide should be 16 times of $2\pi$, and then the total phase shift can achieve $2\pi$. According to the relationship between the propagating constant and phase shift, it is found that this condition is satisfied at around 6.3 GHz. In a similar way, $l = 2$ appears at about 6.6 GHz. Furthermore, when the phase provided by the spoof SPP waveguide is less than the phase lag introduced by the circular metal patches, a reversal in the phase rotation turns up. At about 5.8 GHz, the phase provided by the spoof SPP waveguide is 14 times of $2\pi$, leading to the OAM mode number $l = -1$. Similarly, the situation with $l = -2$ can be reached at around 5.5 GHz. Table 1 provides the precise values of frequencies corresponding to the relevant OAM modes, as well as the calculated value of relevant parameters in the intermediate process. From the comparison between the theoretical arithmetic and full-wave simulation, the accuracy of the analysis can be corroborated due to the tiny deviation of frequency. The cause of these subtle errors can be blamed to the approximate evaluation of the propagation constant, which is inevitable.

The far field radiation patterns at the predicted frequencies are also calculated through MATLAB, serving as a preliminary evidence of the proposed principle and design. The calculated results are given in Figure S1 in Supporting Information.

### Table 1. Different parameter values of the relevant frequencies for each OAM mode.

| Index of OAM mode | $l = -2$ | $l = -1$ | $l = 0$ | $l = 1$ | $l = 2$ |
|-------------------|----------|----------|---------|---------|---------|
| Propagating constant $k$ | 169.6 | 181.58 | 194.72 | 209.5 | 220 |
| Waveguide wavelength $\lambda_g$ (mm) | 37 | 34.6 | 32.2 | 30 | 28.57 |
| Phase change caused by spoof SPP ($2\pi$) | 13 | 14 | 15 | 16 | 17 |
| Phase lag caused by circular patches ($2\pi$) | 15 | 15 | 15 | 15 | |
| Total phase change ($2\pi$) | -2 | -1 | 0 | 1 | 2 |
| Calculated frequency (GHz) (Theoretical arithmetic) | 5.41 | 5.71 | 6.00 | 6.28 | 6.54 |
| Calculated frequency (GHz) (Full-wave simulation) | 5.45 | 5.78 | 6.03 | 6.31 | 6.59 |

2.2. Simulations and Measurements

To illustrate the radiation efficiency, the proposed structure is simulated and fabricated for measurements. All materials in experiments are the same as those in simulations. Scattering (S) parameters (i.e., reflection coefficients $S_{11}$ and transmission coefficients $S_{21}$) are measured using Agilent vector network analyzer (VNA), as shown in Figure 5. With reference to Figure 5a, we observe that the reflection coefficients are less than -10 dB in the whole radiation frequency range, indicating good impedance match. The only difference appears at the minimum values.
between the two results, mainly resulting from system error. The system error cannot be fully eliminated due to the limitation during calibration. Furthermore, the simulation is completed under ideal condition, while the measurement is operated in practical condition, where both the temperature and the vibration of structure have impact on the measured results. Also, the operating frequency of the proposed structure ranges from about 5 GHz to 7 GHz, in which the measured and simulated results agree very well with each other. This can verify the accuracy of the design. From Figure 5b, we clearly see the resonant frequencies corresponding to the relevant OAM modes. Good agreements between simulated and measured results and the comparison in Table 1 firmly validate the proposed theory.

However, for a two-port antenna, such as the antenna proposed here or the leaky-wave antennas which can be easily found in open literatures, low $S_{11}$ and $S_{21}$ can only prove that most of the power is sent to the antenna system. Power could be emitted by the antenna or be absorbed by the materials in the antenna. The latter often leads to the low antenna gain and efficiency. To prove the good performance of the proposed antenna, not only $S_{11}$ and $S_{21}$, but also the antenna efficiency is measured, as shown in Table S1 in Supporting Information. From the low $S_{11}$ and $S_{21}$ as well as the high efficiency, we can conclude that most of the energy is radiated to the free space, which reveals the feasibility of our design. Besides, the gains of the proposed structure at the corresponding frequencies are given in Table S1 in Supporting Information.

In order to further verify the above analysis, a near-field experiment was performed, as shown in Figure 6. The comparison between the simulated and measured normalized near-field distributions is presented in Figure 7, in which Figures 7a–e show the simulated amplitudes, Figures 7f–j are the measured ones, while Figures 7k–o give the simulated phase distributions, and Figures 7p–t are the measured ones. The commercial software, CST Microwave Studio, was used to simulate the designed structure. Due to the limitation of the computer resources, the observation plane is set at 500 mm (10 times of the wavelength at the central frequency) above the proposed structure with an area of $370 \times 370 \text{ mm}^2$. The near-field measurements were completed in an anechoic chamber by using near-field antenna test system, which is composed of a fixed platform and a position controllable probe connected to the Agilent VNA. One of the two 50Ω-coaxial cables acts as the connection between VNA and the sample. The other connects the probe and VNA. The vertical probe located in front of the sample with a distance of 600 mm (12 times of the wavelength at the central frequency) serves as receiver. During the measurement, the probe moves along the $x$- and $y$-directions step by step, and the phase distributions are plotted by the measured data via MATLAB. The whole observation plane in
The mode number of OAM beam is conspicuous. We can get $l = -2, -1, 0, 1, 2$ at 5.5, 5.8, 6.0, 6.3, and 6.6 GHz, respectively, corresponding to the forecasts from the analysis.

From the simulated and measured amplitude distributions of the proposed structure, we can conceive the far-field radiation pattern. However, to get the precise cognition, the far-field radiation patterns are also measured for further verification to the vortex-beam generation (see Figure 8). Figures 8a–e and 8f–j illustrate the simulated and measured normalized three-dimensional (3D) radiation patterns, respectively. At 6 GHz, an ordinary beam is obtained since the mode number is zero, as has been predicted before. Hollow radiation patterns are achieved at other frequencies as expected, except for some imperfect amplitude performance. The main reason for this issue owns to the continuous leak of EM waves, resulting in the unequal radiation amplitudes from different metal patches. For clearance, the intensity along the circular patch is simulated, as shown in Figure S2 in Supporting Information. Both simulated and measured results prove that the previous analysis is accurate and the proposed method to generate vortex beams by spoof SPPs is effective.

Last but not least, the modal content is also an indicator of the characteristic of the vortex beam emitter. To illustrate the mode purity of the emitted beam, the normalized modal content at each resonant frequency is calculated through both simulated and measured phase data, as shown in Figure 9. From this figure, it is clearly observed that at certain frequencies, the expected modes appear to these points, and the other modes are in very small amounts, implying little effects on the final emitted beam. The good agreement between simulated and measured results can be regarded as another success of the proposed design.
3. Conclusion

In summary, we proposed a novel and simple method to generate vortex beams by spoof SPP. The transmission route of EM waves was formed by a looped double-layer spoof SPP waveguide. The beam emitting was realized through a series of circular metal patches. Such circular patches are used not only for the radiation units, but also for resonators to produce the phase shift. The proposed method was verified by both numerical calculation and experiments, and the measured results agree very well with numerical simulations as well as theoretical predictions. Vortex beams with different OAM modes were observed at different frequencies. The mode numbers of $l = -2, -1, 0, 1, 2$ appear at 5.5, 5.8, 6.0, 6.3, and 6.6 GHz, respectively. The normalized modal content at each resonant frequency is also calculated through both simulated and measured results. The high purity of the expected mode can be observed clearly, implying the success of the proposed design. The simple structure and design procedure make the proposed method promising in the future wireless communication systems. Compared with previous methods, this smaller structure possesses more flexible OAM modes, and is much easier to be integrated into systems and ICs. This work is not only an expansion to the application of spoof SPP, but also a simplifier of the vortex-beam generation.

Supporting Information

Supporting Information is available from the Wiley Online Library or from the author.

Acknowledgements

This work was supported in part from the National Natural Science Foundation of China (61631007, 61571117, 61302018, 61401089, 61501112, 61501117), and in part from the 111 Project (111-2-05).

Conflict of Interest

The authors declare no conflict of interest.
Keywords
orbital angular momentum, spoof surface plasmon polaritons, vortex beam

Received: December 2, 2016
Revised: December 18, 2017
Published online: January 15, 2018

[1] L. Allen, M. W. Beijersbergen, R. Spreeuw, J. Woerdman, Phys. Rev. A 1992, 45, 8185.
[2] S. Franke-Arnold, L. Allen, M. Padgett, Laser Photonics Rev. 2008, 2, 299.
[3] D. G. Grier, Nature 2003, 424, 810.
[4] M. Padgett, R. Bowman, Nat. Photonics 2011, 5, 343.
[5] J. E. Curtis, D. G. Grier, Phys. Rev. Lett. 2003, 90, 133901.
[6] A. O’neil, I. MacVicar, L. Allen, M. Padgett, Phys. Rev. Lett. 2002, 88, 053601.
[7] M. Chen, M. Mazilu, Y. Arita, E. M. Wright, K. Dholakia, Opt. Lett. 2013, 38, 4919.
[8] V. Skarka, N. Aleksić, V. Berezhiann, Phys. Lett. A 2003, 319, 317.
[9] S. Bretschneider, C. Eggeling, S. W. Hell, Phys. Rev. Lett. 2007, 98, 218103.
[10] A. Mair, A. Vaziri, G. Weihs, A. Zeilinger, Nature 2001, 412, 313.
[11] R. Fickler, R. Lapkiewicz, W. N. Plick, M. Krenn, C. Schaeff, S. Ramelow, A. Zeilinger, Science 2012, 338, 640.
[12] S. Yu, Opt. Exp. 2015, 23, 3075.
[13] S. Oemrawsingh, A. Aiello, E. Elieel, G. Nienhuis, J. Woerdman, Phys. Rev. Lett. 2004, 92, 217901.
[14] S. Oemrawsingh, X. Ma, D. Voigt, A. Aiello, E. Elieel, J. Woerdman, Phys. Rev. Lett. 2005, 95, 240501.
[15] Z. Dutton, J. Ruostekoski, Phys. Rev. Lett. 2004, 93, 193602.
[16] B. Thidé, H. Then, J. Sjöholm, K. Palmer, J. Bergman, T. Carozzi, Y. N. Istomin, N. Ibragimov, R. Khamitova, Phys. Rev. Lett. 2007, 99, 087701.
[17] F. Tamburini, E. Mari, A. Sponselli, B. Thidé, A. Bianchini, R. Romanato, New J. Phys. 2012, 14, 033001.
[18] Y. Yan, G. Xie, M. P. J. Lavery, H. Huang, N. Ahmed, C. Bao, Y. Ren, Y. Cao, L. Li, Z. Zhao, A. F. Molish, M. Tur, M. J. Padgett, A. E. Willner, Nat. Commun. 2014, 5.
[19] E. Brasselet, M. Malinauskas, A. Žukauskas, S. Juodkazis, Appl. Phys. Lett. 2010, 97, 211108.
[20] G. Tumbull, D. Robertson, G. Smith, L. Allen, M. Padgett, Optical Angular Momentum 2003, 186.
[21] P. Schemmel, S. Maccalli, G. Pisano, B. Maffei, M. W. R. Ng, Opt. Lett. 2014, 39, 626.
[22] C. Zhang, L. Ma, Sci. Rep. 2016, 6, 31921.
[23] A. Tennant, B. Allen, Electron. Lett. 2012, 48, 1.
[24] Q. Bai, A. Tennant, B. Allen, Electron. Lett. 2014, 50, 1.
[25] X. Gao, S. Huang, Y. Song, S. Li, Y. Wei, J. Zhou, X. Zheng, H. Zhang, W. Gu, Opt. Lett. 2014, 39, 2652.
[26] Z.-G. Guo, G.-M. Yang, IEEE Antennas Wireless Propag. Lett. 2016, 99, 1.
[27] K. Liu, H. Liu, Y. Qin, Y. Cheng, S. Wang, X. Li, H.-Q. Wang, IEEE Trans. Antenn. Propag. 2016, 64, 3850.
[28] Y. Yang, W. Wang, P. Moitra, I. I. Kravchenko, D. P. Briggs, J. Valentine, Nano Lett. 2014, 14, 1394.
[29] W. Wang, Y. Li, Z. Guo, R. Li, J. Zhang, A. Zhang, S. Qu, J. Opt. 2015, 17, 045102.
[30] J. Sun, X. Wang, T. Xu, Z. A. Kudyshev, A. N. Cartwright, N. M. Litchinitser, Nano Lett. 2014, 14, 2726.
[31] L. Cheng, W. Hong, Z.-C. Hao, Sci. Rep. 2014, 4, 4814.
[32] E. Karimi, S. A. Schulz, I. De Leon, H. Qassim, J. Upham, R. W. Boyd, Light Sci. Appl. 2014, 3, e167.
[33] N. Yu, P. Genevet, M. A. Kats, F. Aieta, J. P. Tetienne, F. Capasso, Z. Gaburro Science 2011, 334, 333.
[34] P. Genevet, N. Yu, F. Aieta, J. Lin, M. A. Kats, R. Blanchard, M. O. Scully, F. Capasso, Appl. Phys. Lett. 2012, 100, 013101.
[35] P. Genevet, J. Lin, M. A. Kats, F. Capasso Nat. Commun. 2012, 3, 1278.
[36] F. Gambini, P. Velha, C. Oton, S. Faralli, IEEE Photonics Tech. Lett. 2016, 28, 21.
[37] X. Cai, J. Wang, M. J. Strain, B. Johnson-Morris, J. Zhu, M. Sorel, J. L. O’Brien, M. G. Thompson, S. Yu, Science 2012, 338, 363.
[38] R. Li, X. Feng, D. Zhang, K. Cui, F. Liu, Y. Huang, IEEE Photonics J. 2014, 6, 1.
[39] Y. Chen, S. Zheng, H. Chi, X. Jin, X. Zhang, in EuRAD, Paris, France, Sep. 2015.
[40] J. Y. Yin, J. Ren, H. C. Zhang, B. C. Pan, T. J. Cui, Sci. Rep. 2015, 5, 8165.
[41] J. Y. Yin, J. Ren, H. C. Zhang, Q. Zhang, T. J. Cui, Sci. Rep. 2016, 6, 24605.
[42] J. J. Xu, J. Y. Yin, H. C. Zhang, T. J. Cui, Sci. Rep. 2016, 6, 22692.
[43] J. Y. Yin, H. C. Zhang, Y. Fan, T. J. Cui, IEEE Antennas Wireless Propag. Lett. 2016, 15, 865.
[44] B. C. Pan, J. Zhao, Z. Liao, H. C. Zhang, T. J. Cui, Sci. Rep. 2016, 6, 1.
[45] J. Y. Yin, J. Ren, Q. Zhang, H. C. Zhang, Y. Q. Liu, Y. B. Li, X. Wan, T. J. Cui, IEEE Trans. Antenn. Propag. 2016, 65, 1.