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2-bit dual-polarized reconfigurable intelligent metasurface with low power consumption for 6G near-field communication*

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Abstract: Near-field communication using large-scale antenna arrays is one of the hot research topics in 6G communication. Reconfigurable intelligent surface (RIS) is a cost-effective method for manipulating electromagnetic waves in the near field. This article proposes a 2-bit dual-polarized RIS that has the merits of being low cost and having low power consumption, high phase accuracy, and polarization diversity. Each element consists of an aperture-coupled microstrip patch, two SP4T switches, and two groups of microstrip delay lines. 2-bit phase shift is achieved by using only one SP4T switch that controls the connection of four parallel delay branches. Dual polarization is generated by placing two orthogonal slots with two 2-bit phase shifters. A 15×15 RIS prototype operating in the 3.6 GHz band is fabricated and measured. The beam can be scanned in the ±60° range, with peak aperture efficiency of 40.1% for vertical polarization and 38.3% for horizontal polarization. What is more, the total power consumption of the RIS is merely about 100 mW, which is very attractive for massive deployment in 6G near-field communication.

Key words: Reconfigurable intelligent surface (RIS); Dual polarization; Near-field communication; 6G communication
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1 Introduction

Near-field communication is the new emerging technique in 6G communication, providing a higher transmission rate, lower latency, and more reliable connections. The propagation characteristics of electromagnetic waves can no longer be simply regarded as plane waves but rather as spherical waves. This introduces new electromagnetic effects such as beam splitting, polarization diversity, and evanescent waves (Li XR et al., 2022; Wang ZL et al., 2023; Zhang X and Zhang HY, 2023). Many traditional communication algorithms experience performance degradation in 6G near-field scenarios or are unable to fully utilize new features (Jiang H et al., 2023; Alexandropoulos et al., 2024; Dai ZY et al., 2024; Xu JQ et al., 2024).

The reconfigurable intelligent surface (RIS) has great potential in near-field communication because of its ability to proactively change the wireless channel environment (Wymeersch et al., 2020; Mei et al., 2022; Jiang YH et al., 2023). The waves emitted by the base station can be reprocessed by the RIS, such as beam redirection, focusing, and splitting, to enhance the signal quality, fill the fading zone, and suppress interference. Fig. 1 shows the typical RIS-aided scenario for near-field and far-field hybrid communication. The hardware of RIS is applicable

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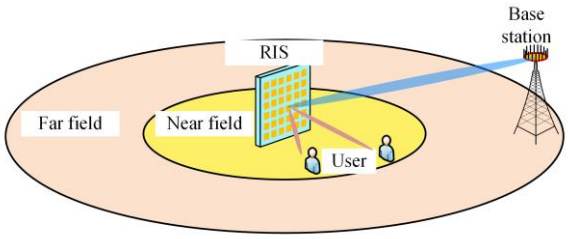

Fig. 1. Typical RIS-aided scenarios for near-field and far-field hybrid communication.

for both near-field and far-field communications. However, there are two significant differences between the two scenarios. Firstly, the array scale of the RIS is much larger in near-field communication, in order to expand the near-field coverage region. Secondly, the beamforming algorithm is different, where the EM wave is regarded as a plane wave in far-field communication and a spherical wave in near-field communication, respectively. Due to the increase of RIS aperture, the range of near-field communication is significantly expanded, and terminal users are within the near-field region. For example, the near-field range of a RIS array with 20×20 half-wavelength-spacing elements operating at 3.6 GHz is about 22 m, and it will further increase as the array scale increases. Dense deployment of RIS is required for wide spatial coverage. Therefore, the cost and power consumption of RIS are crucial factors that will determine its massive application in the future 6G near-field communication.

The passive metasurface is low cost and consumes no power, but suffers from low flexibility due to structural solidification. A RIS loaded with tunable components that can generate an agile beam is more attractive. P-i-n diodes (Dai LL et al., 2020; Luyen et al., 2020; Lee et al., 2022; Zhang QS et al., 2022), varactor diodes (Hum et al., 2007; Perruisseau-Carrier, 2010; Pan et al., 2013), MEMS switches (Perruisseau-Carrier and Skrivervik, 2008; Moghadas et al., 2015), liquid crystal (Li XY et al., 2021; Kim et al., 2023), and graphene (Dash et al., 2022) are widely used tunable components. Among them, the p-i-n diode and MEMS switch can change the phase of cells discretely, while the varactor diode, liquid crystal, and graphene can achieve continuous phase variation at the cost of a complex biasing circuit. MEMS switches have the advantages of low insertion loss and high cut-off frequency. However, the cost

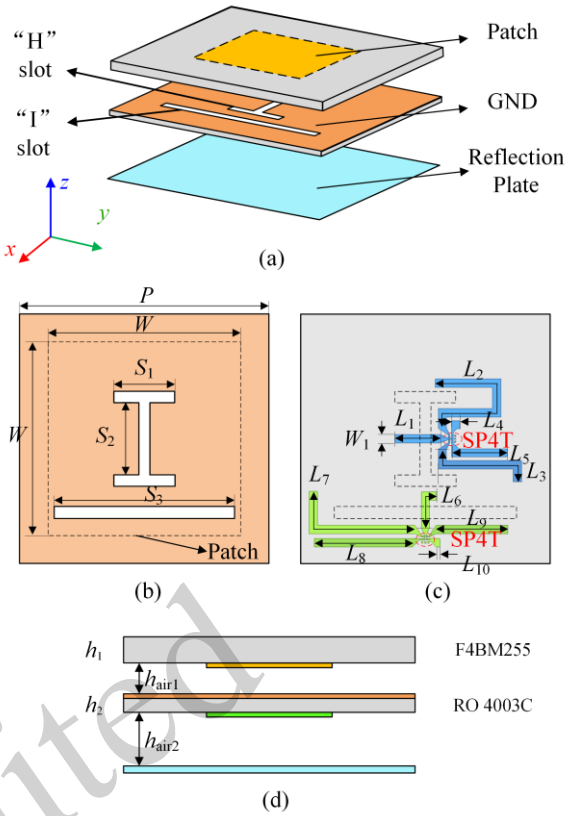

Fig. 2. Structure of the proposed 2-bit dual-polarized element. (a) 3-D exploded view, (b) Ground（GND）layer, (c) transmission line layer, (d) side view of the element.

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and response time still have room for improvement. P-i-n diodes make a good trade-off between cost and performance, where many commercial components are available. In addition, discrete phase quantization can approach continuous phase variation by increasing the number of quantizations. According to the analysis of (Yang et al. 2017), the phase quantization loss of 1-, 2-, and 3-bit phase discretization is 3, 0.6, and 0.2 dB, respectively. Numerous excellent RISs based on p-i-n diodes have been presented in the last decade (Han et al., 2019; Xu HJ et al., 2020; Xiang et al., 2022; Li P et al., 2023; Wu et al., 2023). The drawback of the p-i-n diode is its high power consumption since it is current-driven. The typical biasing power of a p-i-n diode is about 0.8 W. The total power consumption of a RIS will be very large, since it usually includes hundreds of p-i-n diodes.

In this paper, a novel 2-bit dual-polarized RIS operating in the 3.6 GHz band is presented that has low power consumption. Instead of using p-i-n diodes, an SP4T switch is adopted to build the 2-bit phase

shifter. The CMOS fabrication process gives the switch a very low biasing current. The total power consumption of the 15×15 RIS is about 100 mW. To the best of the authors' knowledge, it is the first dual-polarized RIS that uses only one tunable component to achieve a 2-bit phase shift.

2 Design of 2-bit dual-polarized element

This section explains in detail the geometry, operating principle, and key parameters of the 2-bit dual-polarized element.

2.1 Structure of the proposed element

Fig. 2 shows the geometry and dimensions of the proposed 2-bit dual-polarized element, which is based on an aperture-coupled microstrip patch antenna. It includes two substrate layers and one metallic plate. An air gap with a height of 5 mm is inserted into the two substrate layers. The distance between the lower substrate layer and the metallic plate is set as 8 mm. The upper and lower substrate layers are made of F4BM255 (ε_r=2.55, tanδ=0.0015) and RO4003C (ε_r=3.55, tan δ=0.0027), respectively.

The radiating patch is printed on the bottom of the upper substrate layer. It has a square shape to support dual polarizations. The ground plane beneath the patch is printed on the top of the lower substrate layer. An H-shaped slot and an I-shaped slot are etched on the ground plane. The two slots are orthogonally placed to provide dual-polarized coupling. The microstrip transmission lines are printed on the bottom of the lower substrate layers, which can be divided into two groups. Each group has one common microstrip line and four microstrip delay branches. The common line is beneath the H-shaped or I-shaped slot to achieve electromagnetic energy coupling. The four branches have different lengths to achieve different phase delays. The connection between the common line and the four branches is controlled by an SP4T switch. By properly setting the biasing voltages, the switch can connect the common line with one of the four branches.

The center frequency point of the RIS is set as 3.6 GHz, and the operation frequency band is one of the popular Sub-6 GHz bands. The period of the element is 33 mm, corresponding to 0.4 wavelength at 3.6 GHz in free space. All the key parameters are optimized and listed in Table 1.

Table 1 Dimensions of the proposed element (Unit: mm)

parameter	*value*	*parameter*	*value*
L_1	6	L_6	5.75
L_2	24.6	L_7	18.5
L_3	15.3	L_8	13
L_4	1.1	L_9	9.5
L_5	7.4	L_{10}	0.5
S_1	8	P	33
S_2	9.5	W	25.5
S_3	24	W_1	1.1
h_1	1	h_2	0.508

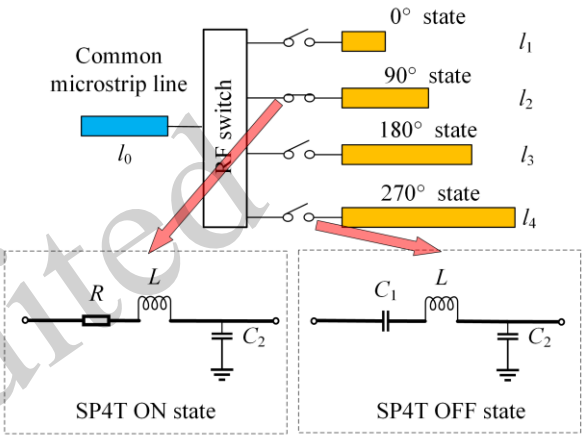

Fig. 3. Equivalent circuit of the SP4T switch.

2.2 Phase tuning with the SP4T switch

As shown in the element structure, the tunable components are SP4T switches rather than the conventional p-i-n diodes. After an investigation of the literature, it is found that the single-pole-multi-throw switch is rarely used in RIS design. In order to understand the characteristics of the SP4T switch, the equivalent circuit of the switch is built, and an evaluation circuit board is fabricated and measured to extract the equivalent parameters (Vilenskiy et al., 2020; Wang EH et al., 2024).

Fig. 3 shows the equivalent circuit of the SP4T switch in the ON and OFF states. In the ON state, the SP4T switch is modeled as a series circuit of a resistor R and an inductor L, while in the OFF state, it is modeled as a series circuit of a capacitor C_1 and an inductor L. There is a grounding capacitor C_2 in both

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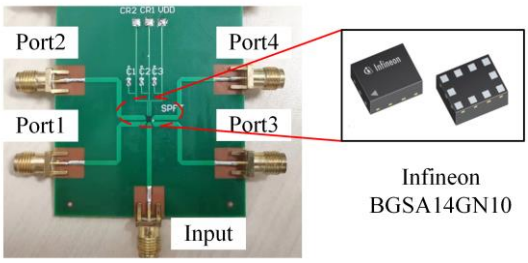

Fig. 4. Evaluation circuit board for the parameter extraction of the SP4T switch.

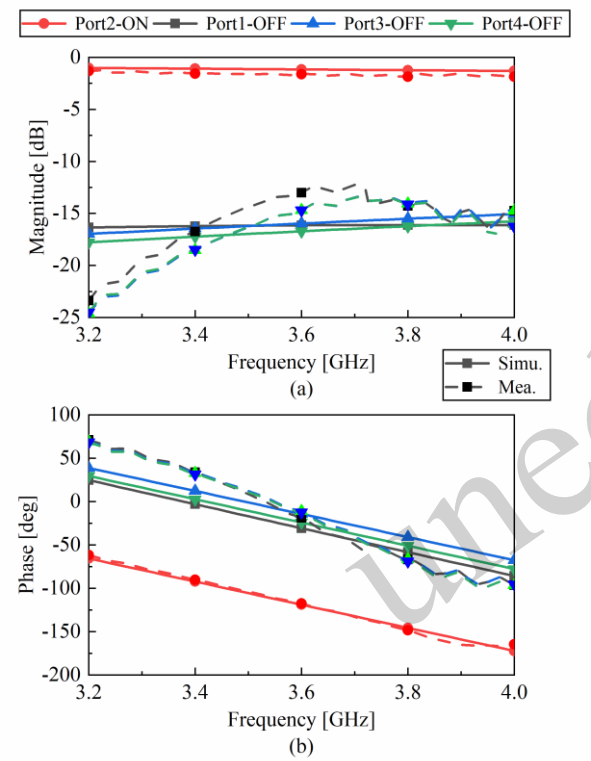

Fig. 5. Simulated and measured transmission coefficients of the evaluating board. (a) Magnitude, (b) phase.

states. The main difference between the ON and OFF states is whether a resistor or capacitor is used.

In order to extract the equivalent parameters of the SP4T switch, an evaluation circuit board is fabricated, as shown in Fig. 4. Considering the operating frequency, the Infineon BGSA14GN10 SP4T switch is adopted, which has the merits of being low cost and having low power consumption. The circuit board has one signal input port and four signal output ports. Two biasing voltage pins are used to control the states of the SP4T switch. By setting high and low voltages on the two pins, the input port is connected with one output port, while the other three output ports are disconnected. The biasing circuit is simple since only binary voltage control is needed.

By setting the bias voltages, Port 2 is connected, while the other three output ports are disconnected. Fig. 5 compares the simulated and measured results. The measured loss in the ON state of the SP4T switch is about 1-dB across the relevant 3.4–3.8 GHz band. In the simulation, the equivalent components are set as follows: R=1.6Ω, C_1=120fF, L=0.1nH, and C_2=42fF. It is shown that the simulated and measured results have good consistency, indicating the feasibility of the equivalent circuit.

Based on the SP4T switch, the required 0°, 90°, 180°, and 270° states for the construction of the 2-bit phase shifter can be generated. The basic model of the element is an aperture-coupled patch with a microstrip delay line. Electromagnetic wave received by the patch will first be transmitted to the microstrip line through the coupling slot. Then, it will be reflected back at the end of the microstrip line, which is open-circuited. Finally, the reflected wave will be coupled to the patch and re-radiated to the free space.

The reflection phase of the element is mainly determined by the length of the microstrip delay line. According to microwave theory, the input impedance of an open-ended transmission line can be calculated using the following equation (Yin et al., 2024).

$$Z_{in}(l)=-jZ_0\times\cot(\beta l) \quad (1)$$

where Z_0 is the characteristic impedance of the transmission line, β is the phase constant at 3.6 GHz, and l is the distance from the open-circuited end to the input port.

From this equation, it is known that any reflection phase could be achieved by simply changing the length of the transmission line. In order to achieve a dynamical phase change, switches are loaded on the transmission line. Considering the phase tuning accuracy and controlling complexity, a 2-bit phase shifter is preferred, which is constructed by deploying four parallel branches with different lengths; 0°, 90°, 180°, and 270° reflection phases can be achieved by properly selecting the lengths of the four delay lines and controlling the working state of the SP4T switch.

2.3 Element performance

The proposed 2-bit dual-polarized element is

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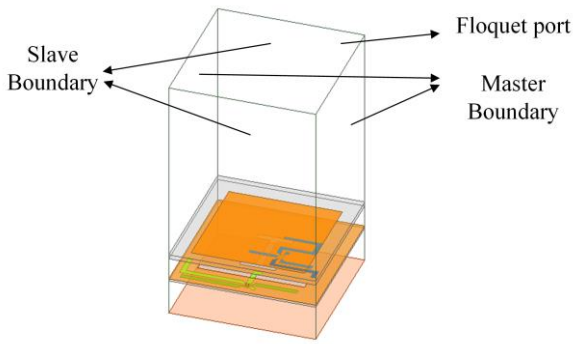

Fig. 6. Simulation setup of the periodic element.

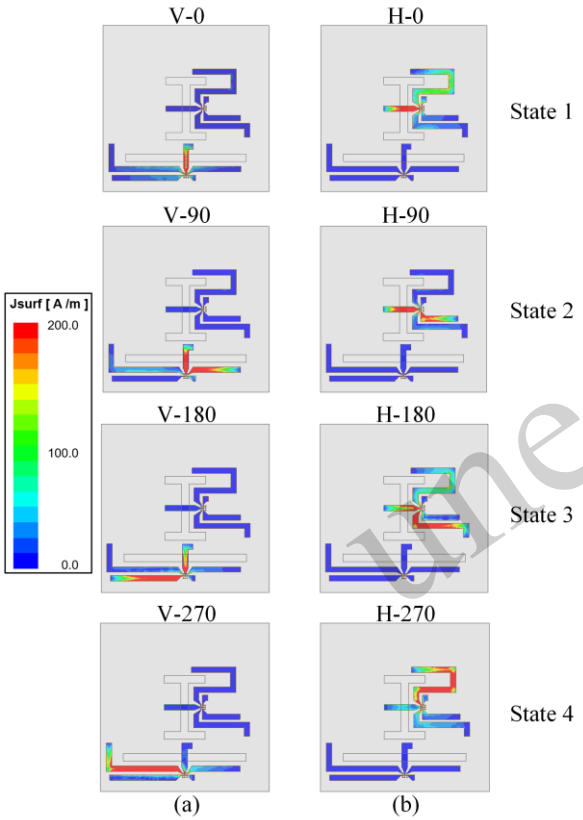

Fig. 7. Average surface current density of the four states on the microstrip lines. (a) Vertical polarization, (b) horizontal polarization.

simulated by using an infinite periodic boundary combined with Floquet port excitation in Ansoft HFS software. Fig. 6 shows the simulation setup of the periodic element. The SP4T switch is modeled as lumped components based on its equivalent circuit. Both transverse electric (TE) mode and transverse magnetic (TM) mode are excited to simulate the incident wave illumination with dual polarizations (Hao et al., 2024).

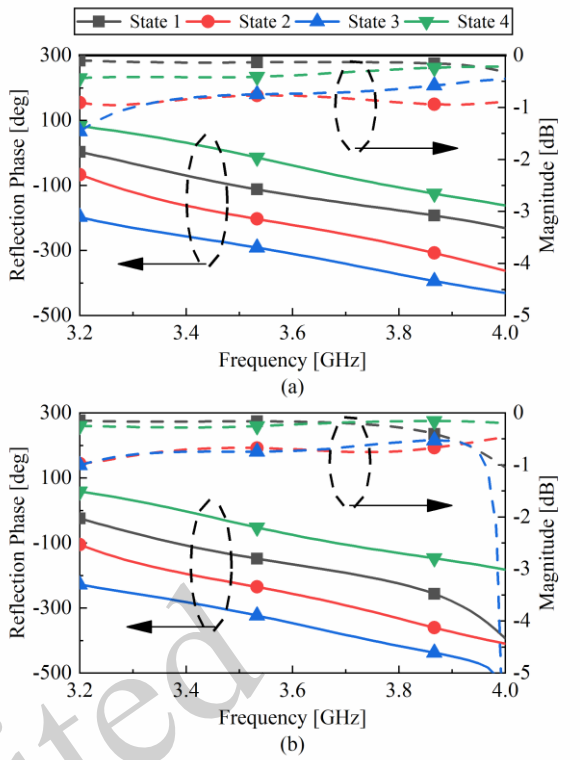

Fig. 8. Reflection phase and magnitude of the 2-bit dual-polarized element. (a) Vertical polarization, (b) horizontal polarization.

In order to explain the operating mechanism of the 2-bit dual-polarized element, the average current distributions on the two groups of the microstrip lines are presented in Fig. 7. Each group has one SP4T switch that controls the connection between the common line and the four parallel branches. In the left column of Fig. 7, it can be seen that the current is mainly concentrated on the line group beneath the I-shaped slot, while the intensity on the line group beneath the H-shaped slot is very weak in all four states. This means that the vertical polarization is excited independently. Furthermore, the current on the connected branch is much stronger than that on the three disconnected branches. Considering that the four branches have different lengths, four different reflection phase states are generated. By properly adjusting the lengths, the four states can provide 0°, 90°, 180°, and 270° phase shifts, corresponding to a 2-bit phase shifter. Similarly, in the right column of Fig. 7, the current distribution depicts that horizontal polarization with a 2-bit phase shifter can also be excited independently. Therefore, the 2-bit dual-polarized element is constructed.

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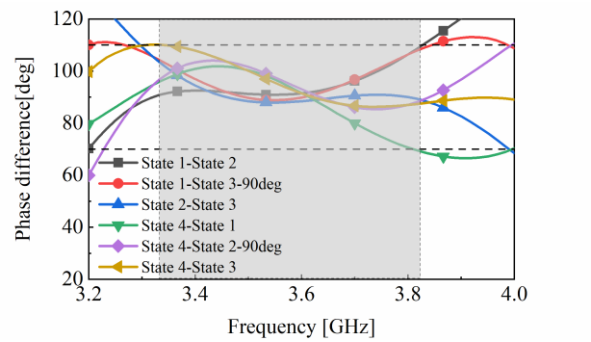

Fig. 9. Phase difference between any two adjacent states with vertical polarization.

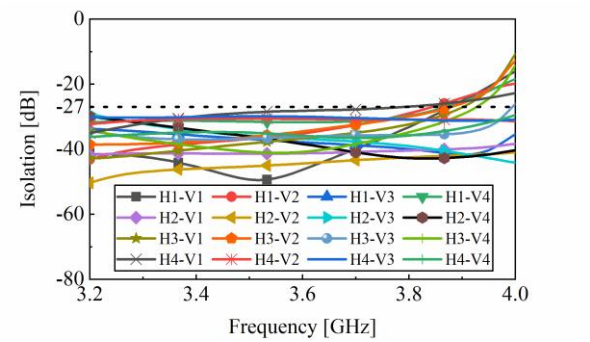

Fig. 11. The isolation between any two states between the vertical and horizontal polarizations.

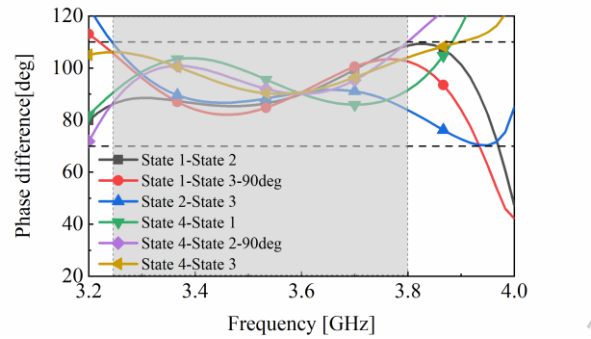

Fig. 10. Phase difference between any two adjacent states with horizontal polarization.

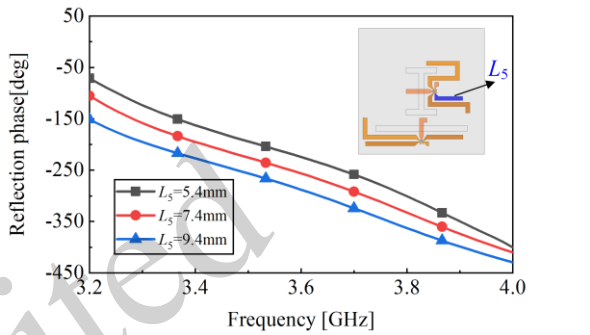

Fig. 12. The reflection phase of State 2 in horizontal polarization with different values of L_5.

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Fig. 8 shows the reflection phase curves and reflection magnitude curves of the two polarizations in the four switching states. The phase curves have good linearity among the four states in both vertical and horizontal polarizations. The reflection amplitude of the two polarizations is generally above -1 dB across the whole bandwidth in all the states, indicating low reflection loss. It is noted that the reflection phase and loss change dramatically at 4 GHz, which is caused by the resonance of the microstrip delay line. However, since it is beyond the band concerned, this deterioration is acceptable in practical applications.

The bandwidth of the proposed 2-bit element is evaluated by using the criterion of a 90°±20° phase difference between any two adjacent phase states. Fig. 9 and Fig. 10 show the calculated results of vertical and horizontal polarizations, respectively. It is observed that the vertical polarization and horizontal polarization have bandwidths of 500 MHz and 550 MHz, respectively, corresponding to 13.8% and 15.3% relative bandwidth. It is noted that the overlapping bandwidth covers the 3.4–3.8 GHz band.

As well as observing the current distribution in Fig. 7, the isolation between the vertical and horizontal polarizations can be further evaluated quantitatively from the coupling between any two states. As shown in Fig. 11, the simulated isolation is always greater than 27 dB within the operating frequency band, no matter which two states of the two polarizations are selected. This verifies that the proposed 2-bit element has a dual-polarization feature.

2.4 Parameter analysis

In order to optimize the performances effectively, the key parameters of the element are studied. There are two types of parameters. The first type only affects a single polarization, such as the dimensions of the microstrip lines and the coupling slots. The second type affects both vertical and horizontal polarizations, such as the dimensions of the radiating patch and the thickness of the air gaps.

The length of the microstrip branches is the key parameter in the first type. Take L_5, for example; this value mainly affects State 2 in the horizontal polarization. Fig. 12 shows the influence of L_5 on the reflection phase of State 2. As L_5 increases, the phase

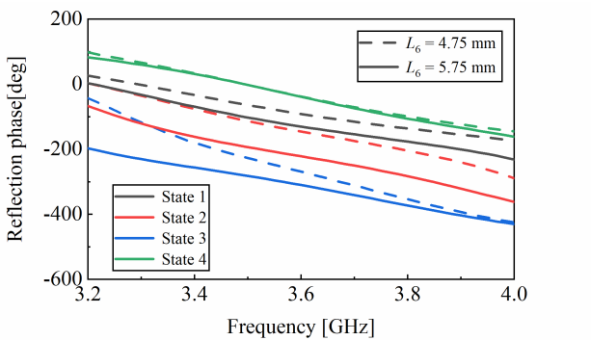

Fig. 13. The reflection phase of the four states in vertical polarization with different values of L_6.

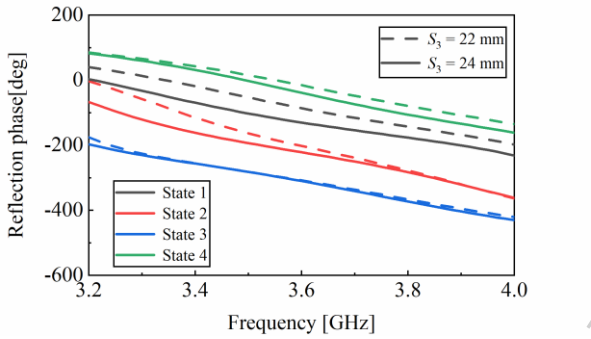

Fig. 14. The reflection phase of the four states in vertical polarization with different values of S_3.

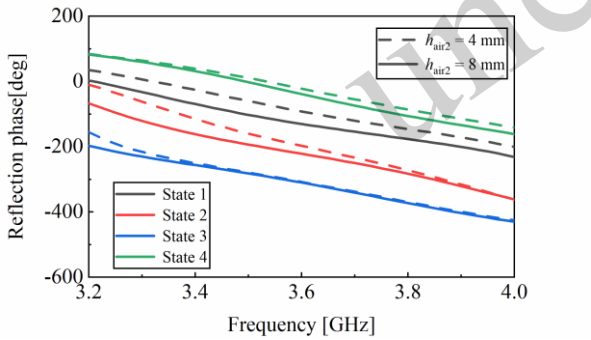

Fig. 15. The reflection phase of the four states in vertical polarization with different values of H_{air2}.

curve shifts down with good linearity in the operating frequency bandwidth. The performances of other states are little influenced, which is not shown for brevity. This implies that this type of parameter can be used to change the phase response of a single state.

The length of the common microstrip line is another key parameter in the first type. Take L_6, for example; it can affect all four states in the horizontal polarization. Fig. 13 shows the influence of L_6 on the reflection phase of the vertical polarization. It has

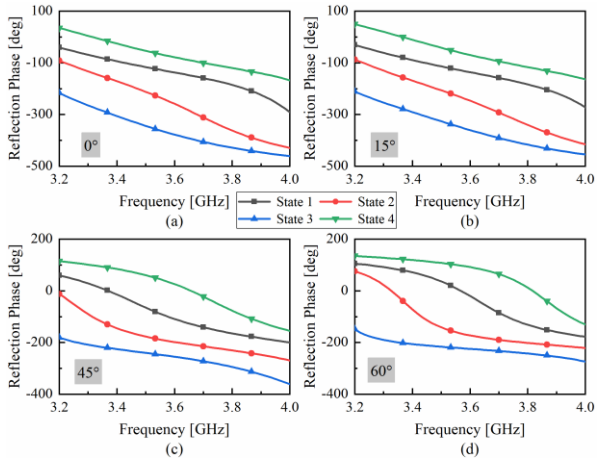

Fig. 16. The reflection phase of the four states in vertical polarization at different oblique incident angles. (a) 0°, (b) 15°, (c) 45°, (d) 60°.

different influences on the four states. To be specific, State 2 and State 3 have large phase shifts when L_6 changes, while State 1 and State 4 are less sensitive. Similarly, the length of the coupling I-shaped slot (S_3) can also affect all four states. As shown in Fig. 14, State 1 and State 2 have large phase shifts when S_3 changes, while State 3 and State 4 are stable. Therefore, the two parameters are effective in optimizing multiple states of the vertical polarization.

The thickness of the air gap is the key parameter in the second type. Fig. 15 shows the influence of h_{air2}, namely the distance between the lower substrate layer and the reflection plate, on the reflection phase. As h_{air2} increases, the linearity among the four curves improves, which means that the bandwidth is wider with a thicker air gap. This parameter can be used to optimize the bandwidth.

From the parametric study, it is known that the proposed design is very flexible in optimizing the performances of the 2-bit element.

Considering that the RIS could be illuminated from different incident angles, the angle sensitivity of the element is carried out. Fig. 16 shows the states of the horizontal polarization with different incident angles. It is seen that the linearity of the curves worsens when the incident angle moves away from 30°, which is a compromise among the multiple angles.

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3 Design and measurement of the RIS

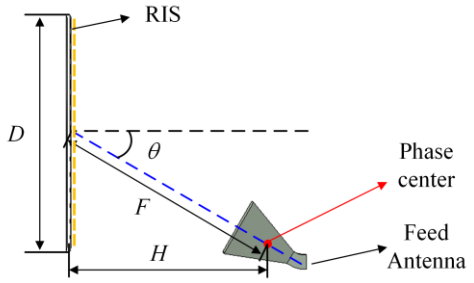

Fig. 17. Simulation model of the 15×15 RIS fed by a horn antenna.

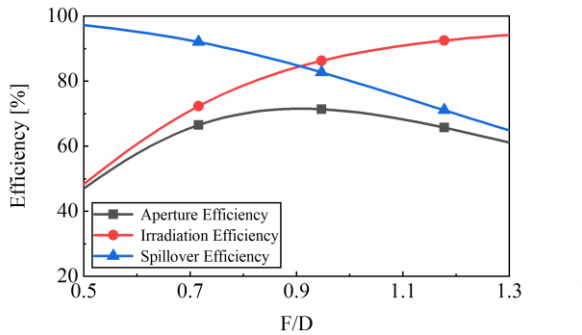

Fig. 18. Calculated aperture efficiency of the RIS with different F/D ratios.

A 15×15 2-bit dual-polarized RIS is designed to verify the beam scanning performances based on the proposed element. Fig. 17 shows the simulation setup of the RIS. A linearly-polarized pyramidal horn antenna with 14.5 dBi gain at 3.6 GHz is employed as the feeder. To reduce the blocking of the reflecting wave, the feed horn is located in the *xoz* plane with an oblique angle θ of 30° to the normal direction.

The feed location is an important parameter to optimize the aperture efficiency. For a horn antenna, the phase center can be easily calculated, which is viewed as the equivalent point source of radiation. The height of the feeder is related to the beamwidth of the horn antenna. Fig. 18 shows the influence of the focal length to diameter (F/D) ratio on efficiency. When the F/D ratio increases, the irradiation efficiency increases, but the spillover efficiency decreases. Taking the two effects into consideration, the aperture efficiency is maximized if the F/D ratio is set as 0.9.

The change process of the RIS with beam direction is described as follows. Firstly, set the specific beam direction (θ_0, φ_0) of the RIS as needed. Then, calculate the phase distribution on the aperture of the RIS according to the beam direction. The required phase compensation of each unit cell can then be obtained. Then, discrete the phase of each unit cell by using 2-bit phase quantization. Finally, set the required biasing voltage of each unit cell through the FPGA control circuit board. In this way, the state of each antenna can change to generate different beams.

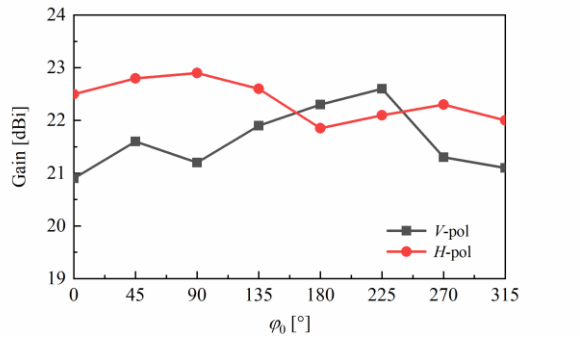

Fig. 19. Broadside gain with different reference phase φ_0.

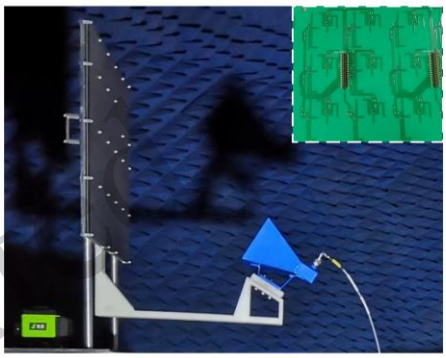

Fig. 20. Photograph of the fabricated RIS prototype in the far-field anechoic chamber.

The reference phase φ_0 of the RIS provides an additional degree of freedom to optimize the gain (Mao et al., 2015; Feng et al., 2018; Li ZP et al., 2021). It can change the phase distribution of the RIS without changing the beam direction. Fig. 19 shows the gain of the broadside beam at 3.6 GHz with the change of φ_0. The gain of the two polarizations can be improved by 1.7 dB and 1 dB by selecting the optimized φ_0.

The RIS prototype is fabricated and measured. Fig. 20 shows the photograph of the fabricated RIS in a far-field microwave anechoic chamber. The two substrate layers are assembled with nylon screws, and the reflection metal plate is made of aluminum sheet.

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The air gaps between the substrate layers and plate are guaranteed by nylon pillars. 3D-printed support is

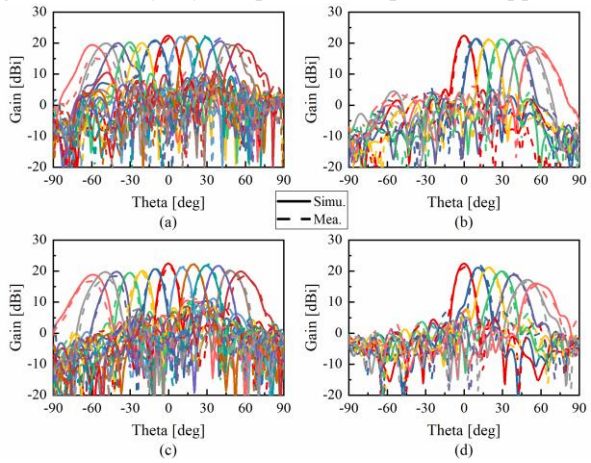

Fig. 21. Simulated and measured radiation patterns of the 2-bit dual-polarized RIS with (a) V-pol in *xoz*-plane, (b) V-pol in *yoz*-plane, (c) H-pol in *xoz*-plane, and (d) H-pol in *yoz*-plane.

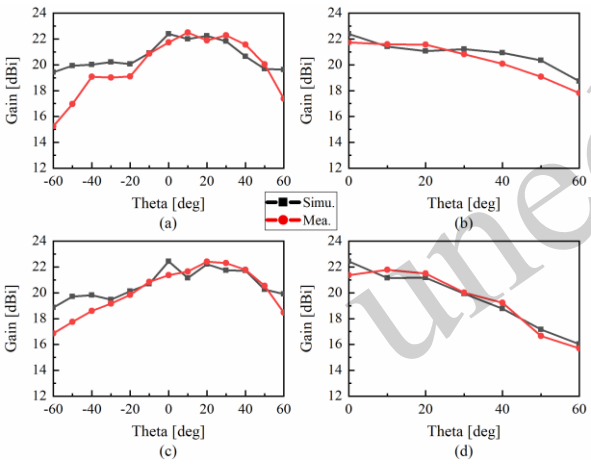

Fig. 22. Peak gain of the 2-bit dual-polarized RIS with (a) V-pol in *xoz*-plane, (b) V-pol in *yoz*-plane, (c) H-pol in *xoz*-plane, and (d) H-pol in *yoz*-plane.

used to fix the feeder on the metal plate.

A low-cost Infineon BGSA14GN10 SP4T switch is adopted, which can cover the 0.1-6 GHz band and only needs two binary biasing voltages. An FPGA circuit board is used to control the SP4T switches, and is installed on the back side of the aluminum sheet. Given that each element includes two SP4T switches, a total of 900 independent biasing voltages are needed for the 15×15 RIS. Therefore, 8-bit shift registers are used to expand the I/O ports of the FPGA board.

Fig. 21 illustrates the measured and simulated radiation patterns at 3.6 GHz. The beam scanning range is from -60° to 60° in the *xoz* plane and from 0°

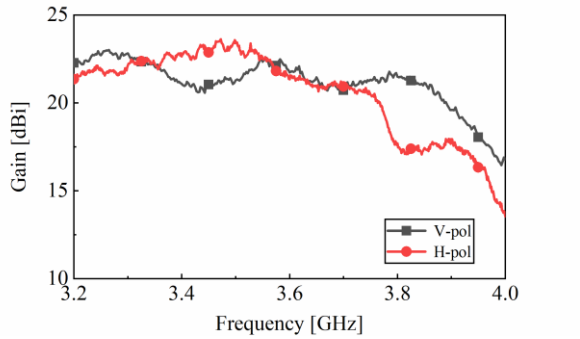

Fig. 23. Measured broadside gain of the two polarizations within the operating frequency band.

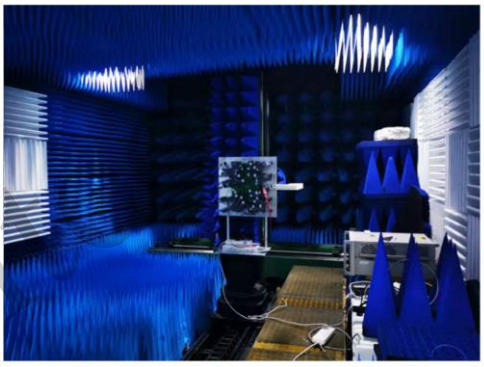

Fig. 24. Measurement setup of the RIS in the near-field anechoic chamber.

to 60° in the *yoz* plane due to symmetry. It is observed that the measured main lobes of the radiation patterns agree well with the simulated results. The measured gain at the scanning angles is slightly lower than the simulated value due to fabrication errors. Fig. 22 shows the simulated and measured gains at different angles. The measured peak gain is 22.5 dBi in the *xoz* plane, corresponding to a peak aperture efficiency of 40.1%.

Fig. 23 shows the measured gain of the broadside beam with frequency variation. The peak gains of the two polarizations reach 23.0 dBi and 23.6 dBi, respectively. In the 3.4–3.8 GHz band, the measured gain is above 20.5 dBi for vertical polarization and is above 17.6 dBi for horizontal polarization.

To further evaluate the performance of the RIS, the prototype is measured in a near-field chamber, as shown in Fig. 24. The amplitude and phase distribution of different scanning beams are measured. Fig. 25 shows the results of the broadside beam. The black

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frame in the center is the outline of the RIS. It can be seen that most of the energy is concentrated on the

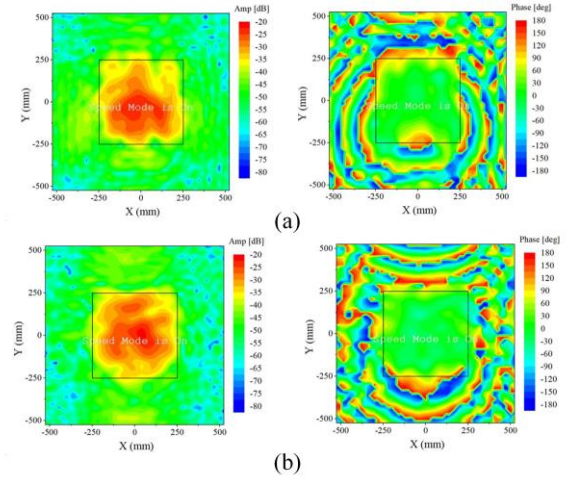

Fig. 25. Measured near-field amplitude and phase distribution of the RIS with broadside beam. (a) Vertical polarization, (b) horizontal polarization.

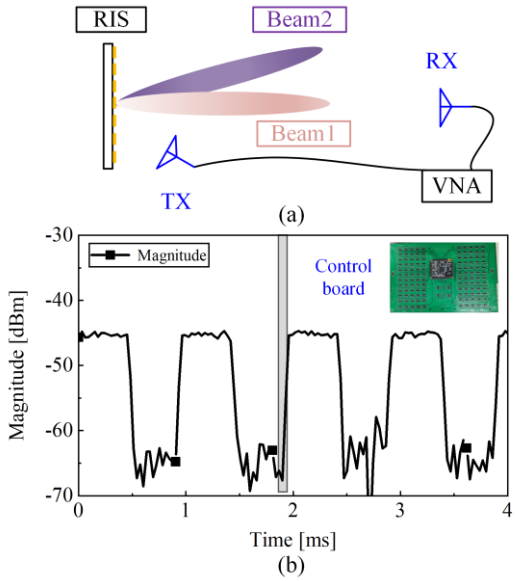

Fig. 26. Beam switching response of the proposed RIS. (a) Measurement setup, (b) response time.

Table 2 Comparisons between the proposed design and other RIS

Ref.	Freq. (GHz)	Elements period (λ_0)	Number of elements	Phase resolution	Polarization	switches per element	Gain (dBi)	AE (%)
Lee et al. (2022)	9.85	0.34	12×12	1-bit	SP	1	17.14	28
Hao et al. (2024)	5.4	0.45	14×14	1-bit	SP	1	21.5	28.5
Xiang et al. (2022)	12.5	0.22	16×16	1-bit	SP	1	20.1	24.0
Dai et al. (2020)	2.3	0.38	16×16	2-bit	AP	5	21.7	31.3
Luyen et al. (2020)	10	0.40	24×24	2-bit	AP	4	25.9	39.4
Wang et al. (2024)	26.1	0.43	24×24	2-bit	SP	2	29.2	27.1
Yin et al. (2024)	3.6	0.40	16×16	2-bit	DP	4	23.1	40.6
This work	**3.6**	**0.40**	**15×15**	**2-bit**	**DP**	**2**	**22.5/ 22.3**	**40.1/ 38.3**

SP: single polarization, AP: agile polarization, and DP: dual-polarization.

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RIS, and the phase fronts are close to each other in the center region.

The RIS response time is also evaluated by measuring the beam switching time. As shown in Fig. 26(a), two beams with different orientations are produced by the RIS—one is pointed to the receiving antenna, and the other is not. By controlling the FPGA board, the beam is switched between the two states. Strong and weak intensities can be observed from the receiving signals, as shown as Fig. 26(b). The rising edge from one state to the other corresponds to the response time. It can be seen that the measured response time is within 0.1 ms.

The total power consumption of the 15×15 RIS is calculated. According to the datasheet of the SP4T switch, the typical supply voltage is 2.8 V, and the supply current is 80 uA; so, the power consumption of a single switch is 0.224 mW. Given that the RIS includes 450 SP4T switches, its total consumption is merely 100.8 mW. The extremely low power consumption characteristic of the RIS is very attractive for large-scale array applications.

Table 2 compares the performances of the proposed RIS with other referenced designs. It can clearly be observed that the proposed RIS can provide 2-bit phase resolution and dual polarization, outper-

forming most of the other designs. Compared to (Yin et al., 2024), there are fewer switches in our design, and the power consumption is much lower.

4 Conclusion

In this article, a 2-bit dual-polarized electronic beam scanning RIS is proposed for 6G near-field applications. The element is based on a dual-polarized slot-coupled patch antenna. Each polarization only needs one SP4T switch to generate 2-bit phase quantization. The two orthogonal polarizations are capable of being controlled independently. A 225-element RIS is fabricated and measured. The main beam can scan from -60° to +60° in both *xoz* and *yoz* planes. The measured peak aperture efficiency achieved for the two polarizations is 40.1% and 38.3%. With the merits of being low cost and having low power consumption, dual-polarization, and 2-bit phase quantization, the proposed RIS has great potential for use in 6G communications.

Contributors

Xiaowei CAO designed the research. Youjia YIN and Yinan HAO processed the data. Changjiang DENG drafted the manuscript. Wei-dong HU, Zhewei FU and Zhiji DENG helped organize the manuscript.

Compliance with ethics guidelines

The authors declare that they have no conflict of interest.

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