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## Article

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# **Full-space multipolarization Janus reconfigurable intelligent metasurface with phase precompensation for bidirectional transmission and asymmetric reflection**

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## **Keywords**

Bidirectional reflection, bidirectional transmission, full-space control, phase precompensation, multipolarization, reconfigurable intelligent metasurface.

## **Abstract**

Reconfigurable intelligent surfaces (RISs) commonly control either reflection or transmission, limiting wavefront manipulation to one half-space or one propagation function. Here we present a full-space multipolarization Janus RIS that supports bidirectional reflection and orthogonally polarized bidirectional transmission within a single aperture. Each meta-atom couples a y-polarized receiving patch to an x-polarized transmitting patch through a microstrip delay line and incorporates two PIN diodes. The 00 and 11 diode states provide 1-bit phase control for forward y-polarized and backward x-polarized reflection, whereas the 01 and 10 states control bidirectional polarization-converting transmission. Delay-line phase precompensation supplies four reflection phases (0°, 90°, 180°, and 270°) and four transmission phases (0°, 45°, 90°, and 135°), mitigating the additional beam produced by 1-bit quantization under plane-wave illumination. Measurements of a 12 × 12 prototype demonstrate beam steering to ±60°, peak gains of 18.31 dBi in reflection and 18.22 dBi in transmission, and measured aperture

efficiencies of 21.3% and 20.7%, respectively. A sparse-aperture configuration further produces four simultaneous beams with orthogonal polarizations in the two half-spaces. The proposed architecture therefore consolidates bidirectional reflection, transmission, polarization conversion, and multibeam operation in one reconfigurable platform.

## Introduction

Sixth-generation wireless systems are expected to connect dense and heterogeneous devices while supporting substantially greater capacity, coverage, and functional integration<sup>1–3</sup>. Meeting these objectives solely through increasingly complex active radio hardware would raise cost and power consumption. Reconfigurable intelligent surfaces (RISs), which reshape the propagation environment through electronically programmable electromagnetic responses, offer a complementary route to improving wireless links<sup>4–8</sup>.

An RIS comprises a two-dimensional array of subwavelength meta-atoms integrated with tunable components such as semiconductor switches<sup>9–11</sup>. By changing the states of these components, the surface reconfigures the amplitude, phase, polarization, or temporal response of an incident wave and thereby converts a passive channel into a controllable propagation environment<sup>12</sup>. This capability has enabled research on integrated sensing<sup>13,14</sup>, wireless power transfer<sup>15,16</sup>, and ambient backscattering<sup>17</sup>, in addition to coverage and capacity enhancement. Most reported RISs nevertheless implement a single dominant function. Representative systems create an auxiliary propagation path<sup>18–23</sup> or steer a focused beam by imposing a spatial phase profile<sup>24–28</sup>, but generally manipulate either reflected<sup>29–34</sup> or transmitted<sup>35–39</sup> waves. The prevalent 1-bit phase resolution simplifies biasing and hardware integration, yet its quantization error can generate an unwanted symmetric beam under plane-wave illumination and reduce the power delivered to the target direction. Higher-bit designs alleviate this effect at the cost of greater circuit complexity, loss, and calibration burden. Additional control of polarization<sup>40</sup>, amplitude<sup>41</sup>, or the time domain<sup>42,43</sup> expands functionality, but reflection-only implementations remain confined to one half-space. Integrated transmitting and reflecting RISs have therefore attracted growing interest<sup>44–46</sup>. Existing implementations typically obtain the two functions by switching them in time<sup>47,48</sup> or partitioning the aperture spatially<sup>49</sup>. These approaches extend coverage to both half-spaces, but the reported 1-bit phase control retains the additional-beam problem under plane-wave excitation. Moreover, many designs optimize only the forward response and do not jointly control the polarization and phase of waves incident from opposite sides.

Phase precompensation provides a hardware-efficient way to suppress the unwanted beam without increasing the number of active phase states. For example, a  $90^\circ$  initial-phase offset between two reflection elements enables single-beam steering with 1-bit tuning<sup>50</sup>, although resonant implementations require multi-objective optimization and are difficult to generalize across several operating modes. Geometric phase offers broader phase coverage through element rotation<sup>51</sup>, but is naturally suited to circular rather than linear polarization. A common phase-precompensation mechanism that operates for linearly polarized bidirectional reflection and transmission has not yet been established. Here we address these limitations with a full-space multipolarization Janus RIS. The receiver-ground-transmitter meta-atom uses two PIN diodes to select between 1-bit reflection and 1-bit polarization-converting transmission, while a microstrip delay line encodes the phase precompensation required for plane-wave operation. Because a reflected wave traverses the delay line twice and a transmitted wave once, the same set of line lengths produces four initial phases for reflection and four corresponding phases for transmission. A  $12 \times 12$  prototype validates bidirectional reflection and transmission, wide-angle beam steering, and a sparse-aperture mode that generates four simultaneous beams. This combination provides full-space wavefront and polarization control without adding active phase states to the meta-atom.

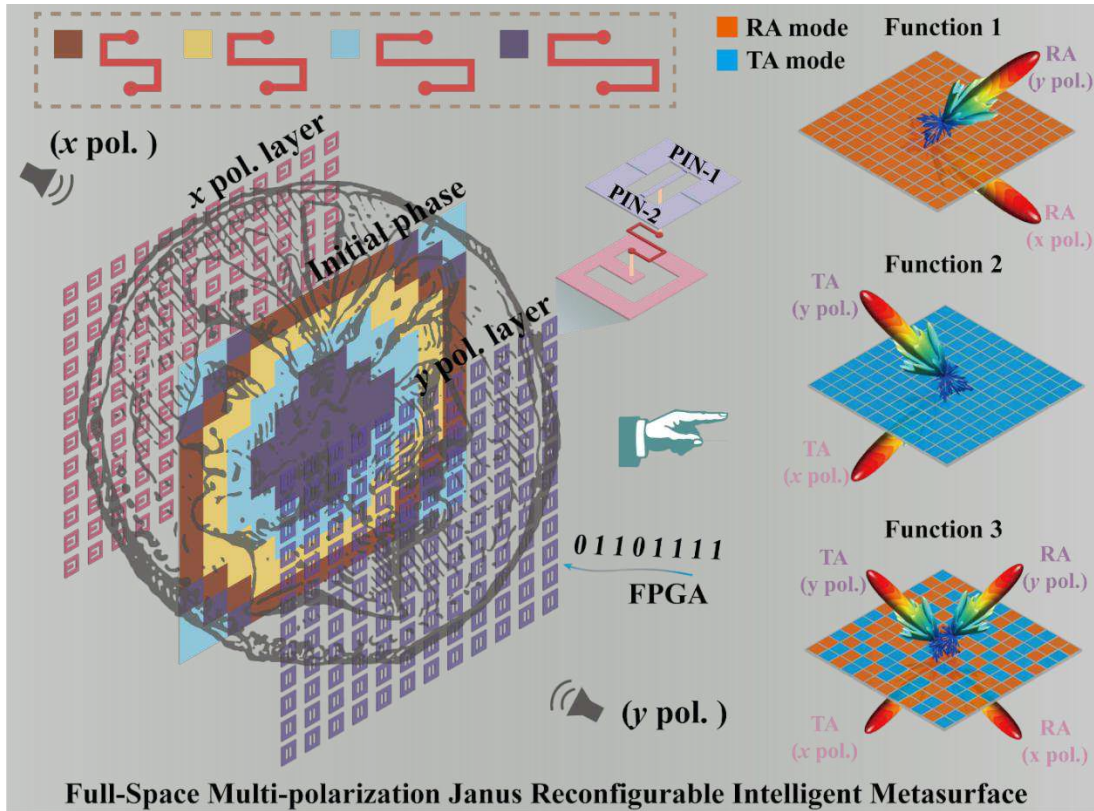


Figure 1. Schematic of the full-space multipolarization Janus RIS.

## Results

Figure 1 summarizes the proposed Janus RIS. The meta-atom comprises a y-polarized receiving patch, a microstrip delay line, and an x-polarized transmitting patch connected by metallized vias. Two PIN diodes on the receiving patch define four bias states. States 00 and 11 produce a  $180^\circ$  phase difference and realize 1-bit reflection control for illumination from either side, corresponding to forward y-polarized and backward x-polarized reflection. States 01 and 10 instead realize 1-bit polarization-converting transmission. The delay line provides a static phase offset in addition to this switchable 1-bit phase. Varying its length yields backward-reflection offsets of  $0^\circ$ ,  $90^\circ$ ,  $180^\circ$ , and  $270^\circ$  and transmission offsets of  $0^\circ$ ,  $45^\circ$ ,  $90^\circ$ , and  $135^\circ$ . The factor-of-two relation follows from the propagation path: a backward-reflected wave traverses the line twice, whereas a transmitted wave traverses it once. For plane-wave illumination, the encoded offsets approximate the phase distribution produced by a near-field horn and mitigate the additional beam associated with 1-bit quantization. We fabricated a  $12 \times 12$  array to evaluate three operating configurations. In reflection mode, near-field forward excitation and approximately plane-wave backward excitation produce simultaneous y- and x-polarized beams. In transmission mode, horns on either side demonstrate bidirectional polarization-converting beam steering. Finally, a genetically optimized 50%-sparse allocation of reflection and transmission states generates four beams concurrently. The following sections establish the meta-atom response, phase-coding method, and measured array performance for each configuration.

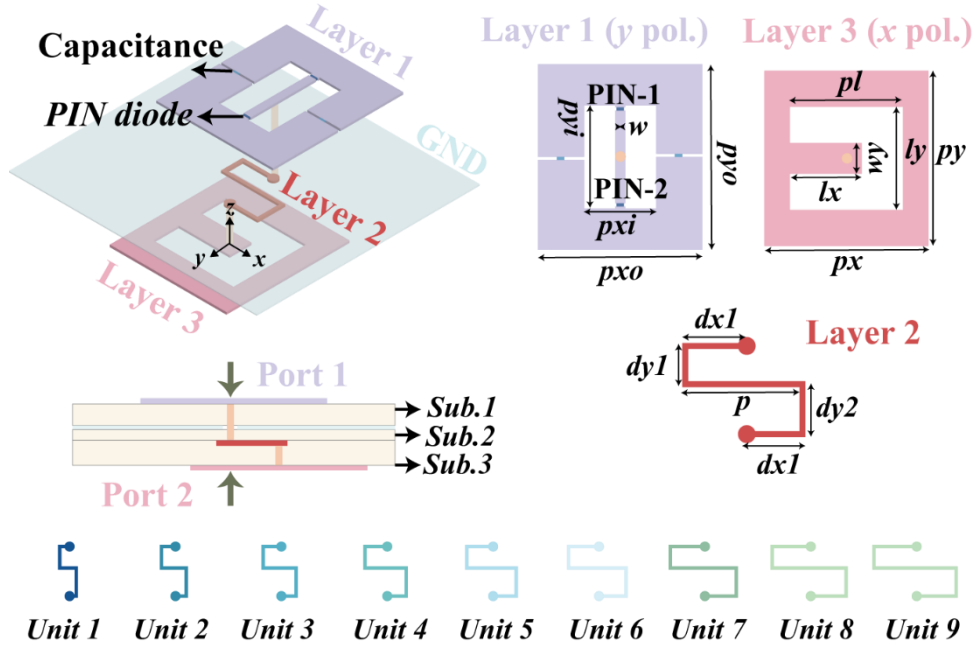


**Figure 2.** Geometry of the meta-atom.

## Meta-atom design

The meta-atom geometry is shown in Fig. 2. It contains three dielectric substrates and four metal layers. Substrates 1 and 3 are 3 mm thick, substrate 2 is 0.254 mm thick, and adjacent substrates are bonded with 0.1 mm Rogers 4450F adhesive films. All three substrates have a relative permittivity of 3.55 and a loss tangent of 0.005. From top to bottom, the metal layers form a y-polarized receiving patch, a ground plane, a meandered microstrip delay line, and an x-polarized transmitting patch. Two metallized blind vias connect the delay line to the patches while remaining isolated from the ground plane. The receiving patch incorporates two SMP1340 PIN diodes, and two 3.5 pF capacitors allow the diodes to be biased independently; the equivalent circuits are given in Supplementary Note 1. States 00 and 11 provide the two reflection phases for both forward y-polarized and backward x-polarized incidence, whereas states 01 and 10 provide the two transmission phases with orthogonal polarization conversion. The meandered line minimizes the footprint required for phase precompensation. Its length is varied through the parameter  $p$  while the blind-via positions remain fixed; Fig. 2 shows  $p$  from 2 to 10 mm in 1 mm steps. For backward reflection, the wave couples from the transmitting patch into the delay line, reflects at the receiving patch, returns through the line, and reradiates from the transmitting patch. Forward reflection occurs directly at the receiving patch and therefore does not acquire this delay-line offset. A transmitted wave traverses the line only once, so its static phase shift is half the corresponding backward-reflection shift. Supplementary Note 1 provides the complete dimensions and operating-state definitions.

Figure 3a presents the reflection magnitude and phase distribution for forward y-polarized reflection mode with  $p=2$  mm. The results demonstrate 1-bit phase modulation between PIN states 00 and 11 across 3.6-3.9 GHz. Both states maintain reflection magnitudes above -0.9 dB, indicating low loss. Magnitude and phase characteristics for forward y-polarized reflection with  $p$  varying from 2-10 mm are shown in Figures S2a-c (supporting information). Thus, forward y-polarized reflection maintains consistent initial phase regardless of delay line length. Figures 3b-d display backward x-polarized reflection characteristics. Varying  $p$  from 2-10 mm shifts the resonant frequency of backward x-polarized reflection. Figure 3b shows reflection coefficients for state 00. Increasing  $p$  initially reduces the resonant frequency, followed by periodic frequency oscillations with further  $p$  increases. State 11 exhibits analogous behavior. Figure 3d presents backward x-polarized reflection phases. Distinct phase curves emerge for different  $p$  values in state 00. At 3.85 GHz, the phases corresponding to 2-10 mm are about -247°, -289°, -333°, -378°, -421°, -459°, -493°, -521°, and -546°, respectively.

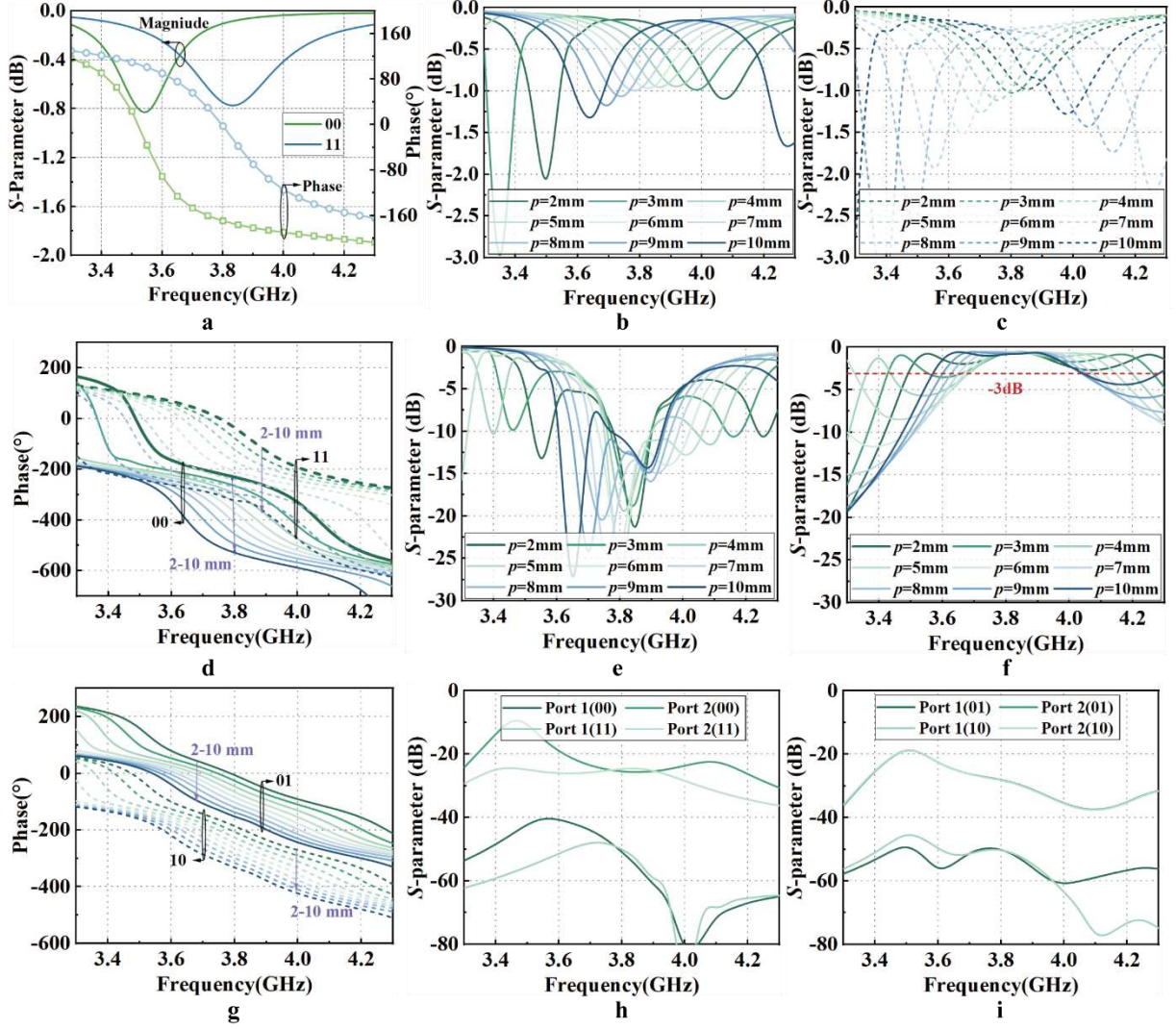


Figure 3. Simulated meta-atom performance. (a) S-parameters for forward y-polarized reflection. Backward x-polarized reflection magnitude in the (b) 00 and (c) 11 states, and (d) reflection phase in both states. Transmission-mode (e) reflection and (f) transmission coefficients in the 10 state, and (g) transmission phase in the 01 and 10 states. Polarization isolation in the (h) 00/11 and (i) 01/10 states.

Using  $p = 2$  mm as the  $0^\circ$  reference, the relative phases of 3-10 mm are  $-42^\circ$ ,  $-86^\circ$ ,  $-131^\circ$ ,  $-174^\circ$ ,  $-212^\circ$ ,  $-246^\circ$ ,  $-274^\circ$  and  $-299^\circ$ , respectively. State 11 shows comparable phase characteristics. We selected  $p = 2/4/6/9$  mm to achieve  $\sim 0^\circ/90^\circ/180^\circ/270^\circ$  reflection initial phases. All  $p$  values maintain 1-bit phase control between states 00 and 11. Figures 3e-g show reflection and transmission coefficients, and transmission phases in transmission mode. Because states 10 and 01 have identical coefficients, only state 10 is shown in Figs. 3e and 3f. State 01 coefficients appear in Figures S3a-b (supporting information). The reflection coefficient for port 2 is given

in Figures S3c-d (supporting information). These results are consistent with the reflection coefficient at port 1. Transmission coefficients exceed -3 dB from 3.7-4 GHz. Reflection coefficients remain below -10 dB across 3.78-3.9 GHz. Figure 3g displays transmission phases for states 01 and 10 with  $p=2-10$  mm. Taking 3.85 GHz as the observation point, it can be seen that the phases corresponding to state 10 when  $p$  is 2-10mm are  $-209^\circ$ ,  $-230^\circ$ ,  $-249^\circ$ ,  $-268^\circ$ ,  $-286^\circ$ ,  $-303^\circ$ ,  $-319^\circ$ ,  $-335^\circ$  and  $-350^\circ$ , respectively. Using  $p = 2$  mm as the  $0^\circ$  reference, the relative phases of 3-10 mm are  $-21^\circ$ ,  $-40^\circ$ ,  $-59^\circ$ ,  $-77^\circ$ ,  $-94^\circ$ ,  $-110^\circ$ ,  $-126^\circ$  and  $-141^\circ$ , respectively. State 01 exhibits similar behavior.  $p=2/4/6/9$  mm were chosen to achieve  $\sim 0^\circ/45^\circ/90^\circ/135^\circ$  transmission phases. These  $p$  values match those for backward x-polarized reflection, ensuring phase design consistency between modes. All  $p$  values maintain 1-bit phase control between states 01 and 10 across the band. Figures 3h-i present cross-polarization characteristics. Figure 3h shows the isolation between x-polarized and y-polarized in port 1 at 00 and 11 states. The corresponding isolation at port 2 is also shown. The cross-polarized response remains below -20 dB from 3.6-4 GHz, confirming excellent polarization purity for both reflection modes. Fig. 3i shows the isolation between x-polarized and y-polarized in port 1 for states 01 and 10. The corresponding isolation at port 2 is also shown. The cross-polarized response is likewise below -20 dB over 3.6-4 GHz. These results confirm high polarization isolation in transmission mode.

### Reconfigurable intelligent metasurface

The far-field pattern of an  $M \times N$  array is

$$E(\theta, \varphi) = \sum_{n=1}^N \sum_{m=1}^M f_{mn}(\theta, \varphi) e^{-j(\varphi(m,n) + kmd \sin \theta \cos \varphi + knd \sin \theta \sin \varphi)} \quad (1)$$

where  $\theta$  and  $\varphi$  denote the elevation and azimuth angles, respectively;  $f(\theta, \varphi)$  is the element factor;  $\varphi(m, n)$  is the phase of element  $(m, n)$ ;  $d$  is the element period; and  $k_0$  is the free-space wavenumber. For a beam directed toward  $(\theta, \varphi)$ , the required continuous phase compensation is

$$\varphi(m, n) = -kmd \sin \theta \cos \varphi - knd \sin \theta \sin \varphi \quad (2)$$

For an array with a prescribed initial phase  $\varphi_0(m, n)$ , the compensation required to direct the beam toward  $(\theta, \varphi)$  becomes

$$\varphi(m, n) = -kmd \sin \theta \cos \varphi - knd \sin \theta \sin \varphi - \varphi_0(m, n) \quad (3)$$

A 1-bit meta-atom provides only two switchable phase states,  $0^\circ$  and  $180^\circ$ . The continuous phase  $\varphi(m, n)$  is therefore quantized as

$$\varphi(m, n) = \begin{cases} 0^\circ & 0 \leq \varphi \leq 180^\circ \\ 180^\circ & 180^\circ \leq \varphi \leq 360^\circ \end{cases} \quad (4)$$

In a reflectarray, a horn in the near field illuminates the aperture with a spherical rather than uniform phase front. The feed-induced phase  $\varphi_f(m, n)$  is

$$\varphi_f = k\sqrt{(x-x_0)^2 + (y-y_0)^2 + (z-z_0)^2} \quad (5)$$

where  $(x, y, z)$  is the position of an array element and  $(x_0, y_0, z_0) = (0, 0, F)$  is the feed position. The focal-distance-to-aperture-size ratio is  $F/D$ , where  $D$  is the array dimension. Under near-field horn illumination, the phase compensation for beam steering is therefore

$$\varphi(m, n) = -kmd \sin \theta \cos \varphi - knd \sin \theta \sin \varphi - \varphi_f(m, n) \quad (6)$$

The resulting phase profile is implemented through 1-bit quantization. Under approximately plane-wave illumination, all  $M \times N$  elements would otherwise have the same  $0^\circ$  initial phase, and binary quantization produces a symmetric additional beam. We instead encode a nonuniform initial phase with the microstrip delay lines. The target offsets are calculated from the path-length difference between a notional near-field horn and each array element using Eq. (5), but are physically implemented in the aperture for plane-wave illumination. The switchable phase state for each element is then selected using Eq. (3).

The meta-atoms were assembled into a  $12 \times 12$  array with a 33 mm period. Each receiving element uses three DC lines: one ground line and two lines that independently bias the PIN diodes. Three 33 nH inductors isolate the RF path from the bias network. Figure 4a shows the fabricated aperture, and Fig. 4b shows the anechoic-chamber measurement arrangement; additional fabrication drawings are provided in Supplementary Fig. S4. As summarized in Fig. 4c, the RIS operates in reflection, transmission, and simultaneous transmission-reflection configurations.

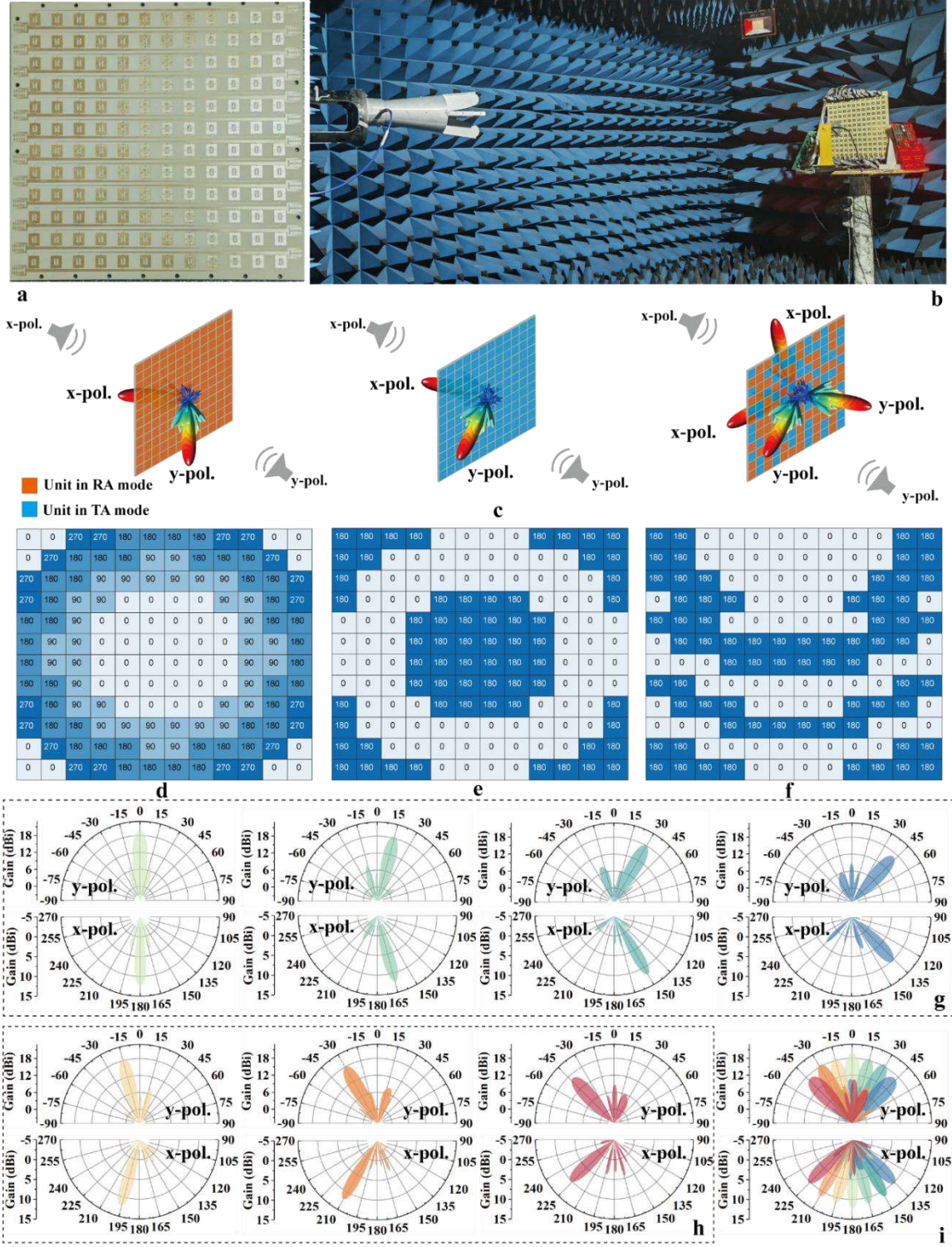


Figure 4. Fabrication and reflection-mode performance. (a) Fabricated RIS. (b) Anechoic-chamber measurement arrangement. (c) RIS operating configurations. (d) Initial-phase distribution for backward x-polarized reflection. Reflection-mode phase codes for steering to (e)  $0^\circ$  and (f)  $30^\circ$ . Bidirectional E-plane reflection for scans toward the (g) positive and (h) negative half-axes at 3.85 GHz. (i) Bidirectional H-plane reflection at 3.85 GHz.

### Simultaneous forward y-polarized and backward x-polarized reflection

In reflection mode, states 00 and 11 provide the  $0^\circ$  and  $180^\circ$  switchable phases for both forward y-polarized and backward x-polarized reflection (Fig. 4c). The two responses therefore operate simultaneously. To obtain a single dominant beam for backward reflection under plane-wave illumination, the microstrip lines implement the initial-phase distribution calculated with Eq. (5). We used  $F/D = 0.8$  for this calculation and placed the horn for forward reflection at the same  $F/D$ . The encoded phase offsets for backward reflection consequently approximate the feed-induced offsets for forward reflection, allowing both beams to use the same binary code when steered to horizontally symmetric directions. The continuous and 2-bit-quantized initial-phase distributions are shown in Supplementary Fig. S5a and Fig. 4d, respectively. The horn-induced phase for forward reflection was quantized as

$$\varphi_f = \begin{cases} 0^\circ & 0^\circ \leq \varphi < 90^\circ \\ 90^\circ & 90^\circ \leq \varphi < 180^\circ \\ 180^\circ & 180^\circ \leq \varphi < 270^\circ \\ 270^\circ & 270^\circ \leq \varphi < 360^\circ \end{cases} \quad (7)$$

With these offsets, a common code steers the forward and backward reflected beams to horizontally symmetric directions. Figures 4e and 4f show the codes for forward steering to  $(0^\circ, 0^\circ)$  and  $(0^\circ, 30^\circ)$ , which correspond to backward directions of  $(0^\circ, 180^\circ)$  and  $(0^\circ, 165^\circ)$ , respectively. Codes for forward steering to  $(0^\circ, 15^\circ)$  and  $(0^\circ, 45^\circ)$  are provided in Supplementary Figs. S5b and S5c. The simulated E-plane patterns in Figs. 4g and 4h confirm bidirectional steering, with y polarization in the forward half-space and x polarization in the backward half-space; the H-plane results in Fig. 4i show the same behavior. Supplementary Fig. S6 gives the corresponding frequency-dependent results. Figures 5a and 5b compare measured E-plane patterns at 3.85 and 3.8 GHz with the simulations, and Figs. 5c and 5d provide the H-plane measurements. The prototype maintains identifiable steered beams up to  $\pm 60^\circ$ . Measurements at 3.9 GHz are reported in Supplementary Fig. S7. Overall, the measured and simulated patterns consistently demonstrate simultaneous forward y-polarized and backward x-polarized reflection.

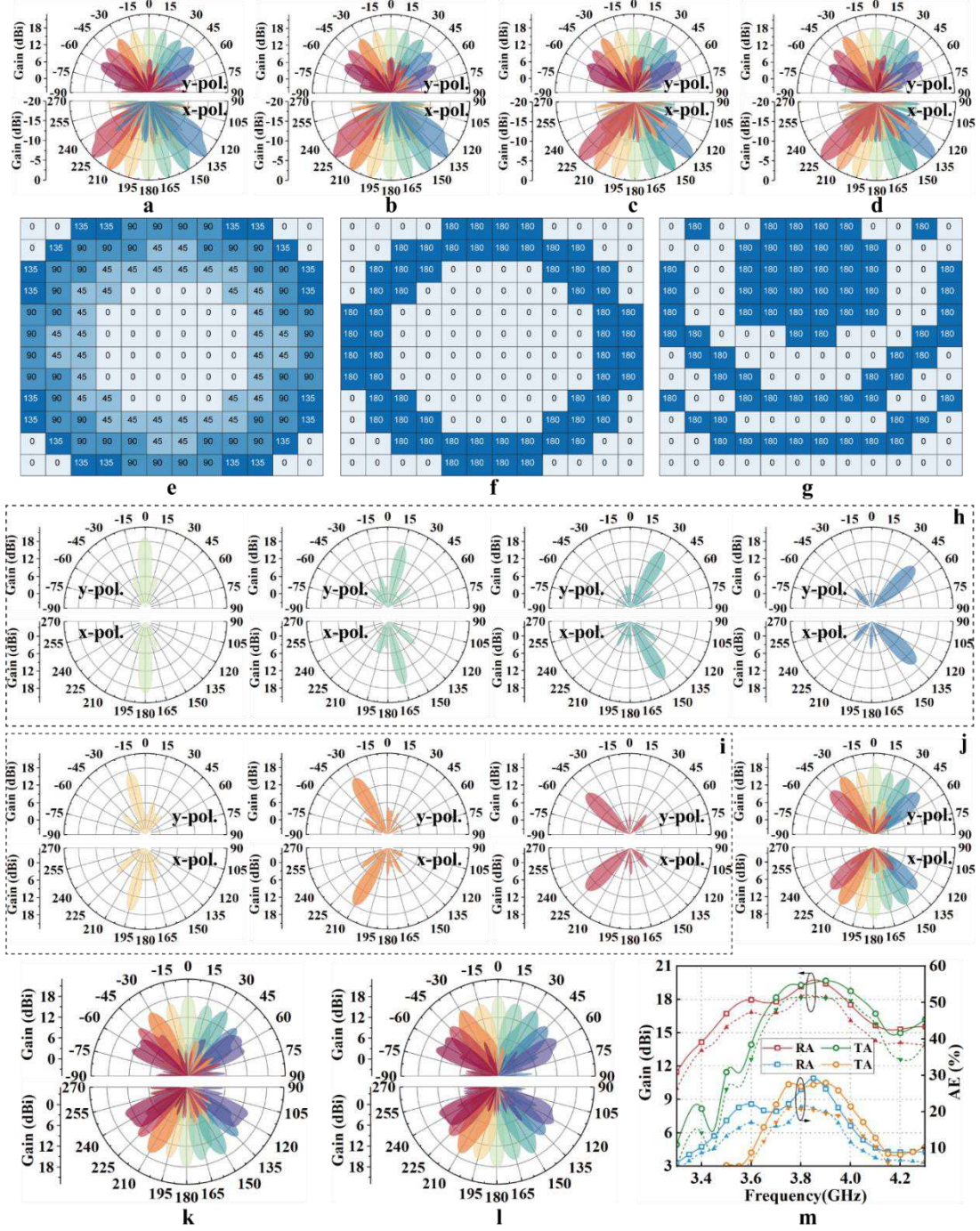


Figure 5. Reflection- and transmission-mode measurements. Measured bidirectional reflection in the E-plane at (a) 3.85 and (b) 3.8 GHz and in the H-plane at (c) 3.85 and (d) 3.8 GHz. (e) Initial-phase distribution for transmission. Transmission-mode phase codes for steering to (f) 0° and (g) 30°. Bidirectional E-plane transmission toward the (h) positive and (i) negative half-axes at 3.85 GHz. (j) Bidirectional H-plane transmission at 3.85 GHz. Measured bidirectional transmission in the (k) E-plane and (l) H-plane at 3.85 GHz. (m) Measured bidirectional gain and aperture efficiency in reflection and transmission.

### Bidirectional x- and y-polarized transmission

Transmission mode uses states 01 and 10, which differ in phase by  $180^\circ$  (Fig. 4c). Forward transmission converts an incident x-polarized wave to a transmitted y-polarized wave, whereas backward transmission converts an incident y-polarized wave to a transmitted x-polarized wave. Because a transmitted wave traverses the microstrip line once, its initial phase is half the backward-reflection phase produced by the same line length (Fig. 5e). These static offsets may also provide a hardware-dependent physical key for communication applications. For the measurements, the near-field horn was again positioned at  $F/D = 0.8$ . The required phase for each transmission element is

$$\varphi(m, n) = -kmd \sin \theta \cos \varphi - knd \sin \theta \sin \varphi - \varphi_0(m, n) - \varphi_f(m, n) \quad (8)$$

The phase distribution is implemented with 1-bit quantization. Figures 5f and 5g show the codes for forward transmission toward  $(0^\circ, 0^\circ)$  and  $(0^\circ, 30^\circ)$ ; the corresponding backward beams point toward  $(0^\circ, 180^\circ)$  and  $(0^\circ, 165^\circ)$ . Codes for forward steering to  $(0^\circ, 15^\circ)$  and  $(0^\circ, 45^\circ)$  are shown in Supplementary Figs. S8a and S8b. The simulated E-plane patterns in Figs. 5h and 5i demonstrate bidirectional transmission, and the H-plane patterns in Fig. 5j show comparable behavior. The forward and backward transmitted beams are horizontally symmetric and have y and x polarization, respectively. Supplementary Fig. S9 provides results at additional frequencies. Measured E- and H-plane patterns at 3.85 GHz (Figs. 5k and 5l) reproduce the simulated steering and remain well defined up to  $\pm 60^\circ$ ; measurements at 3.8 and 3.9 GHz are given in Supplementary Fig. S10. Figure 5m summarizes the gain and aperture efficiency. In reflection, the simulated peak gain is 19.76 dBi at 3.85 GHz and the measured peak is 18.31 dBi at the same frequency; the 1-dB gain bandwidth is 3.76-3.98 GHz. The simulated peak aperture efficiency is 29.2% at 3.85 GHz, whereas the measured maximum is 21.3% at 3.8 GHz. In transmission, the simulated peak gain is 19.68 dBi at 3.9 GHz and the measured peak is 18.22 dBi at 3.85 GHz; the 1-dB gain bandwidth is 3.7-4.05 GHz. The corresponding simulated and measured peak efficiencies are 27.9% at 3.9 GHz and 20.7% at 3.8 GHz.

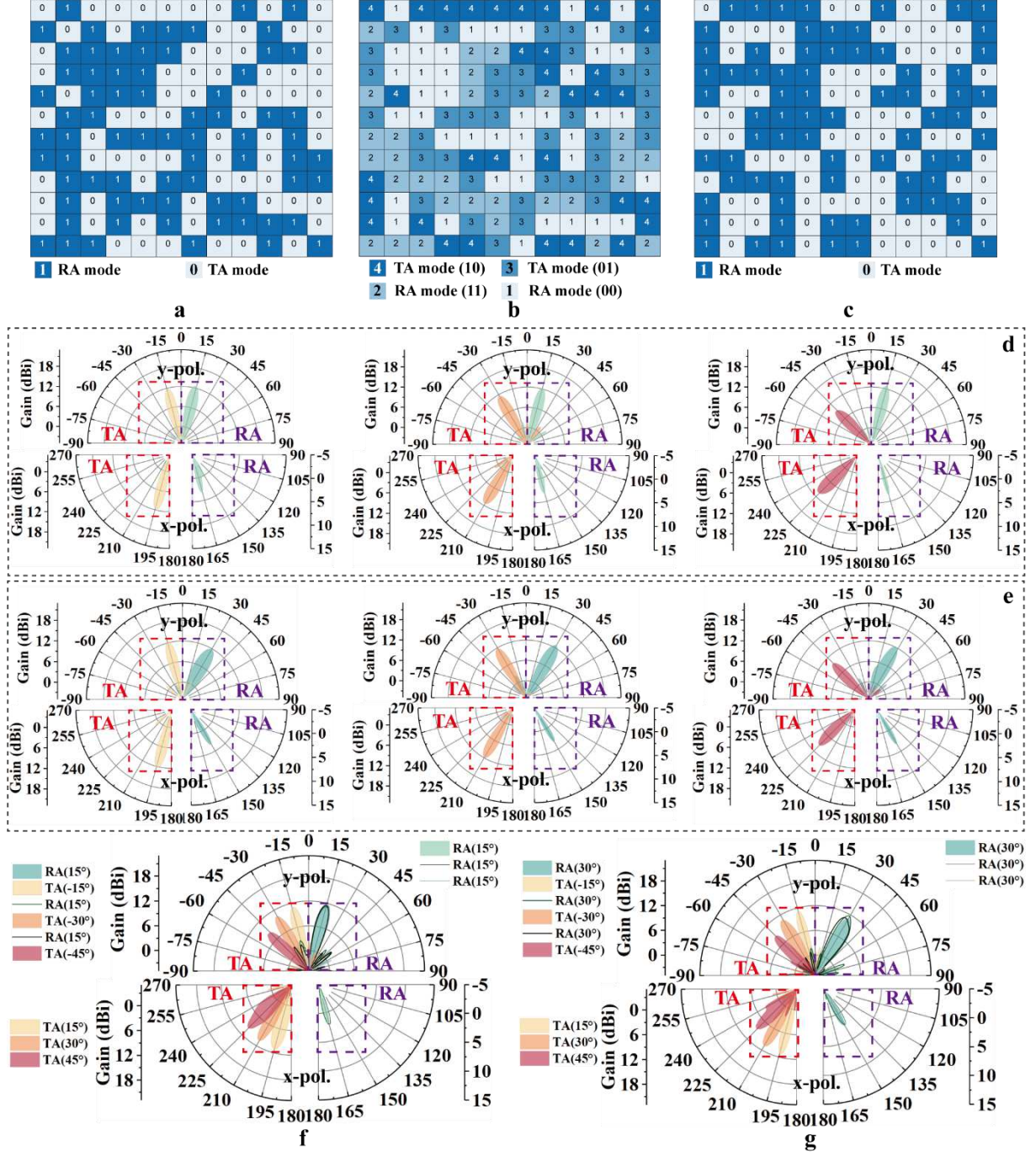


Figure 6. Simultaneous transmission-reflection operation. Optimized array allocations for forward y-polarized reflection at 15° and transmission at (a) -15° and (c) -30°; (b) phase code for the -15° case. Simulated beams for reflection at (d) 15° and (e) 30°, with transmission at -15°, -30°, and -45°. Measured beams for reflection at (f) 15° and (g) 30°, with transmission at -15°, -30°, and -45°.

### Simultaneous forward and backward reflection and transmission

The preceding reflection and transmission results use time-division switching. Concurrent operation instead requires a spatially interleaved aperture (Fig. 4c) in which half of the elements are assigned to each function. We optimized this 50%-sparse allocation with a genetic algorithm. First, a  $12 \times 12$  binary matrix  $A$  was initialized with equal probabilities of 0 and 1, where 1 denotes reflection and 0 denotes transmission; the corresponding excitation masks are  $A$  and  $1 - A$ . Second, target directions were assigned to the reflection and transmission beams, and Eqs. (1)-(4) were used to obtain the binary phase codes. Third, the two far-field patterns were calculated using their respective masks and codes. Finally, candidate layouts were evaluated with the sidelobe-based objective

$$Fitness_{TRA} = Fitness_{RA} + Fitness_{TA} \quad (9)$$

$$Obj = Max(Fitness_{TRA}) \quad (10)$$

$$Fitness_{TRA} = \sum W(\theta, \varphi) \times SLL\_fun(\theta, \varphi) \quad (11)$$

Here  $SLL\_fun(\theta, \varphi)$  is the sidelobe level for a beam directed toward  $(\theta, \varphi)$ , and  $W(\theta, \varphi)$  is the corresponding weight. The fitness is the weighted sum of  $SLL\_fun$  over all prescribed directions. The optimization terminates when the target fitness or the maximum number of generations is reached.

Figure 6a shows the optimized allocation for forward y-polarized reflection at  $15^\circ$  and forward y-polarized transmission at  $-15^\circ$ , and Fig. 6b shows its four-state code. Codes 1 and 2 correspond to reflection states 00 and 11, whereas codes 3 and 4 correspond to transmission states 01 and 10. The same aperture simultaneously produces backward x-polarized reflection at  $165^\circ$  and backward x-polarized transmission at  $195^\circ$ . Figure 6c gives the allocation for reflection at  $15^\circ$  and transmission at  $-30^\circ$ ; additional cases are provided in Supplementary Fig. S11. Simulations in Figs. 6d and 6e confirm four beams when the reflection direction is  $15^\circ$  or  $30^\circ$  and the transmission direction is  $-15^\circ$ ,  $-30^\circ$ , or  $-45^\circ$ . For the representative  $15^\circ/-15^\circ$  case, the forward and backward reflection gains are 12.65 dBi and 3.21 dBi, respectively, and the forward and backward transmission gains are 11.75 dBi and 11.85 dBi. The substantially lower backward-reflection gain reveals a pronounced imbalance among the four beams. Measurements in Figs. 6f and 6g reproduce the four-beam response for both reflection angles and all three transmission angles. The two y-polarized beams occupy the forward half-space and the two x-polarized beams occupy the backward half-space, demonstrating simultaneous full-space multipolarization control.

## Discussion

The proposed Janus RIS combines bidirectional reflection, polarization-converting bidirectional transmission, and simultaneous four-beam operation in a single 1-bit aperture. Its distinguishing feature is the microstrip-line phase precompensation: the same four line lengths provide  $0^\circ/90^\circ/180^\circ/270^\circ$  offsets for backward reflection and  $0^\circ/45^\circ/90^\circ/135^\circ$  offsets for transmission, while the PIN states retain simple binary control. Measurements of the  $12 \times 12$  prototype agree with the simulated steering trends in both principal planes and across 3.8-3.9 GHz. The measured peak gains reach 18.31 dBi in reflection and 18.22 dBi in transmission, with 1-dB gain bandwidths of 3.76-3.98 GHz and 3.7-4.05 GHz, respectively. The simultaneous transmission-reflection configuration extends this functionality to four beams, but spatial partitioning reduces the available aperture for each function; the 3.21 dBi backward-reflection gain in the representative four-beam case illustrates this trade-off. The measured peak aperture efficiencies of 21.3% and 20.7% also leave scope for reducing diode, conductor, dielectric, and feed losses. In addition, the present validation uses a  $12 \times 12$  aperture and asymmetric near-field/approximately plane-wave excitation, rather than a complete communication link. Larger apertures, integrated feeding and control electronics, and link-level measurements will be required to establish scalability and system-level benefit. Within these limits, the results show that phase-precompensated binary meta-atoms can consolidate several full-space wave-control functions without increasing the number of active phase states.

## Methods

**Design and simulation.** The meta-atom was modeled by full-wave electromagnetic simulation using the receiver-ground-transmitter architecture described above. Substrates 1 and 3 were 3 mm thick, substrate 2 was 0.254 mm thick, and adjacent layers were bonded with 0.1 mm Rogers 4450F films. All substrates were assigned a relative permittivity of 3.55 and a loss tangent of 0.005. The receiving patch contained two SMP1340 PIN diodes and two 3.5 pF DC-blocking capacitors; three 33 nH inductors isolated the RF path from the DC bias lines. The diode ON state was represented by a series resistance and inductance ( $R_{on} = 0.85 \Omega$  and  $L_n = 0.45$  nH), and the OFF state by a series capacitance and inductance ( $C_{off} = 0.21$  pF and  $L_n = 0.45$  nH), as detailed in Supplementary Note 1. Array phase codes were calculated with Eqs. (1)-(8), and the simultaneous transmission-reflection element allocation was optimized with the sidelobe objective in Eqs. (9)-(11).

**Fabrication.** A  $12 \times 12$  RIS with a 33 mm element period was fabricated to validate the design. The receiving array, ground plane, microstrip-delay layer, and transmitting array were aligned and bonded as shown in Supplementary Fig. S4 and Fig. 4a. Each element was connected to one ground line and two independently controlled PIN-diode bias lines.

**Measurements.** The prototype was characterized in a microwave anechoic chamber using the arrangement in Fig. 4b. Far-field patterns were recorded in the E- and H-planes at 3.8, 3.85, and 3.9 GHz for forward y-polarized reflection, backward x-polarized reflection, and bidirectional transmission. The horns were positioned at  $F/D = 0.8$  for the near-field measurements. Reflection and transmission patterns were measured for scan angles up to  $\pm 60^\circ$ , and the sparse-aperture configuration was evaluated for simultaneous four-beam operation. Gain and aperture efficiency were measured for the forward reflection and transmission configurations and compared with the simulations in Fig. 5m.

Supplementary information

Supplementary information accompanies this manuscript.

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## Competing interests

The authors declare no competing interests.

## Data availability

The data supporting the findings of this study are available from the corresponding author upon reasonable request.

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