

Article

Single-Varactor Dual-Polarized Reconfigurable Intelligent Surface with Shared-Bias Analog Tuning

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Abstract

Reconfigurable intelligent surfaces (RISs) can modify a propagation environment using electronically controlled reflecting cells. This paper presents a 5.4-GHz dual-polarized RIS, characterized over 5.0–5.5 GHz, in which one shared-bias varactor per cell tunes two orthogonally oriented resonant layers. The simulated co-polarized response provides an analog phase span of approximately 250°. A two-mode Floquet analysis evaluates the polarization-resolved co- and cross-polarized reflection magnitudes and separates cross-polarized leakage from dissipation. Across three representative states, the simulated co-to-cross discrimination is 12.0–13.9 dB, 84.1–96.6% of the incident power is reflected when both polarizations are counted, and absorption is 3.4–15.9%; loss-disabled comparisons identify varactor series resistance as the dominant represented dissipative sensitivity. Unit-cell measurements were also performed in a single-mode WR159 waveguide for three representative capacitance states. A 6×6 prototype was subsequently tested in two proof-of-concept fixed-receiver experiments. Genetic algorithm optimization improved $|S_{21}|$ between two horn antennas, and the single retained SNR trajectory for each horn orientation showed an approximately 2 dB within-run rise. The work demonstrates low-complexity, shared-bias analog field control, and link optimization, while the common control variable does not provide independently selectable TE and TM phase states.

Keywords: reconfigurable intelligent surface; dual polarization; polarization isolation; varactor diode; single active element



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1. Introduction

Wireless networks are being pushed toward higher data rates, lower latency, and denser connectivity by applications such as immersive media, the Internet of Things (IoT), and vehicular communications [1–3]. As systems move from the 5th Generation (5G) toward the 6th Generation (6G) [4–6], maintaining reliable coverage in complex propagation environments has become a central problem [7,8]. Reconfigurable intelligent surfaces (RIS) address this problem by turning the propagation channel itself into a controllable part of the link.

A RIS is a planar array of sub-wavelength unit cells, each able to alter the phase, and sometimes the amplitude, of the wave it reflects. By setting the phase of every cell, the surface can steer, focus, or split a reflected beam without active radio-frequency chains, which makes it an attractive low-cost, low-power way to improve a link [9–12]. The behavior of a RIS is largely fixed by its unit cell: the phase resolution it offers, its reflection

loss, its polarization capability, and the number of active tuning components it needs. Each active component adds DC power, bias wiring, cost, and a potential point of failure, so the number of active elements per cell is a key hardware metric, especially for the large surfaces that practical deployments require [13–15].

Reported reconfigurable surfaces can be grouped by how they realize the per-cell phase shift. Discrete designs use PIN diodes or selected varactor states: Liang et al. [16] demonstrated eight states spanning 315° , Wang et al. [17] realized a wideband 1-bit 20×20 RIS, and Shekhawat et al. [15] used a 1-bit 32×32 tiled aperture with fixed random pre-phasing to suppress quantization lobes. Sayanskiy et al. [18] and Pei et al. [19] reported 400- and 1100-element varactor-loaded 1-bit prototypes, respectively. Continuous designs avoid coarse phase quantization: Li et al. [20] demonstrated a 12×12 liquid-crystal reflectarray, while Kim et al. [21] simulated continuous 300° tuning in two independently tunable frequency bands using stacked, multi-varactor sub-cells. For dual-polarization operation, Yin et al. [22] demonstrated a 2-bit 16×16 reflectarray with independently controlled orthogonal polarizations and 40.56% measured aperture efficiency but required two PIN diodes per polarization.

Recent metasurface research also extends beyond communication-oriented reflection-phase control. Dong et al. [23] integrated speech-command interpretation, target tracking, wireless power transfer, and communication around a digitally coded surface. Niu et al. [24] used a hybrid neural network to inverse-design an optically transparent 1-bit surface for ultrabroadband radar-cross-section reduction, which targets scattering suppression rather than controllable link enhancement. Si et al. [25] developed complex-amplitude Janus control for asymmetric three-dimensional holographic reconstruction, whereas the present work controls the reflected field at a fixed receiver. Dong et al. [26] demonstrated a transparent dual-band rectifying surface whose relevant metric is RF-to-DC conversion efficiency, not reflection-phase coverage. Tang et al. [27] used transmissive space–time-coded meta-atoms for multiplexed vortex-beam generation, in contrast to the present reflective surface with bias states held constant during each link evaluation. These works demonstrate intelligent control, inverse design, energy conversion, holographic multiplexing, and space–time modulation. They are therefore relevant examples of recent metasurface functionality.

Table 1 provides a broad comparison of representative unit-cell architectures, while Tables 2 and 3 focus on recent tunable dual-polarization cells, including Rains et al. [28] and Yurtoglu et al. [29]. The comparison does not show a single architecture dominating every metric. The present work combines one varactor per cell, one shared control variable, and moderate simulated polarization isolation at the evaluated normal-incidence states.

Table 1. Comparison of unit-cell architectures in representative reconfigurable surfaces.

Ref.	Freq. (GHz)	Tuning (No./Cell)	Phase Control	Polarization Control
Li [20]	Ka band	Liquid crystal	Continuous, 360°	Single
Liang [16]	N/R	Varactor (1)	3-bit, 315°	Single
Wang [17]	22.5–29.5	PIN (1)	1-bit, $180^\circ \pm 20^\circ$	Single
Pei [19]	5.8	Varactor (2)	1-bit, $\sim 180^\circ$	Single
Sayanskiy [18]	5.2	Varactor (2)	1-bit, 180°	Single
Shekhawat [15]	27.2	PIN (1)	1-bit, $180^\circ \pm 20^\circ$	Single
Yin [22]	3.6	PIN (4; 2/pol.)	2-bit	Dual, independent
Kim [21]	4.45/12	Varactor (multiple)	Continuous, 300° /band	Dual-band CP
This work	5.4	Varactor (1)	Analog, 250° co-pol. span	Dual linear, shared bias

N/R: not reported in the source material reviewed. CP: circular polarization.

Table 2. Focused architecture comparison of recent tunable dual-polarization unit cells.

Ref.	f_0 (GHz)	Pol.	Control Relation	Tuner (No.)	Phase Range	Cell Size	Array
Rains [28]	3.5	DL	Independent TE/TM	Var. (2)	248°; 1-bit use	$0.35\lambda_0$	48×16
Yin [22]	3.6	DL	Independent TE/TM	PIN (4)	2-bit	N/R	16×16
Kim [21]	4.45/12	DB-CP	Separate bands	Var. (8)	290°/300°	$0.44/0.324\lambda_0$	Row (S)
Yurtoglu [29]	15	DL	Independent TE/TM	Var. (2)	270°	$0.50\lambda_0$	3×3 M; 16×16 S
This work	5.4	DL	Shared TE/TM state pair	Var. (1)	250° co-pol. span	$0.36\lambda_0$	6×6

DL: dual linear; DB-CP: dual-band circular polarization; Var.: varactor; S: simulated; M: measured; N/R: not reported.

Table 3. RF-performance and tuning comparison of recent tunable dual-polarization unit cells.

Ref.	Reflection Loss (dB)	Bandwidth	Bandwidth Definition	Pol. Isolation (dB)	Tuning Range
Rains [28]	≤ 4.2 (M)	160 MHz	1-bit: $180^\circ \pm 20^\circ$	20 (M)	2.5–10 V
Yin [22]	N/R	8.06%	1-dB array gain (M)	N/R	N/R
Kim [21]	$\leq 3.4/6.5$ (S)	N/R	–	N/R	0.04–0.24 pF
Yurtoglu [29]	≤ 4.5 S; ~ 7 M	400 MHz	$\geq 270^\circ$ phase span	61 S; 24 M	0–10 V
This work	0.15–0.75 total; 0.32–1.02 co-pol (S)	54.5 MHz (1.01%) (S)	$\geq 250^\circ$ co-pol.; XPD ≥ 10 dB at 3 states	11.98–13.88 (S)	0–30 V

S: simulated; M: measured; N/R: not reported. For this work, total reflection loss counts both co- and cross-polarized reflected power; the co-polarized entry is attenuation in the original incident basis. The reported range covers three simulated capacitance states at 5.4 GHz and normal incidence. Bandwidth values are not directly interchangeable because the cited studies use different definitions.

For this work, the reported 54.5-MHz (1.01%) simulated bandwidth spans 5.385–5.439 GHz. It is the common interval in the recovered dense co-polarized sweep for which both TE and TM phase spans are at least 250°. Over that interval, the two-mode sweep gives XPD of at least 10.54 dB at the three representative states (1.0, 1.7, and 2.7 pF). Because isolation was not evaluated at every capacitance state defining the dense phase span, this is a qualified combined criterion rather than an all-state full-matrix bandwidth.

The proposed cell stacks two orthogonally oriented resonant layers and uses one varactor to tune their coupled response through a shared DC bias. The main contributions are as follows:

- A single-varactor RIS cell with a shared-bias, capacitance-dependent coupled response for TE- and TM-oriented incident fields, together with a quantitative account of its residual cross-polarized leakage and dissipative-loss sensitivity.
- A compact 12-bit, 0–30 V bias-control board built around a field-programmable gate array (FPGA), digital-to-analog converters (DACs), and operational amplifiers that refreshes the entire 36-cell array synchronously; the DAC and amplifier stages consume a combined 0.78 W under maximum-voltage loading.
- A fabricated 6×6 surface evaluated in proof-of-concept, fixed-receiver $|S_{21}|$, and SNR optimization experiments for both polarization configurations.

Section 2 describes the unit-cell design, operating principle, and simulated response. Section 3 presents the three-setting waveguide comparison, control board, and array assembly. Section 4 reports the two fixed-receiver optimization experiments, and Section 5 concludes the paper.

2. Unit-Cell Design and Simulation

2.1. Unit-Cell Geometry

Figure 1 shows the multilayer unit-cell geometry designed for operation at 5.4 GHz.

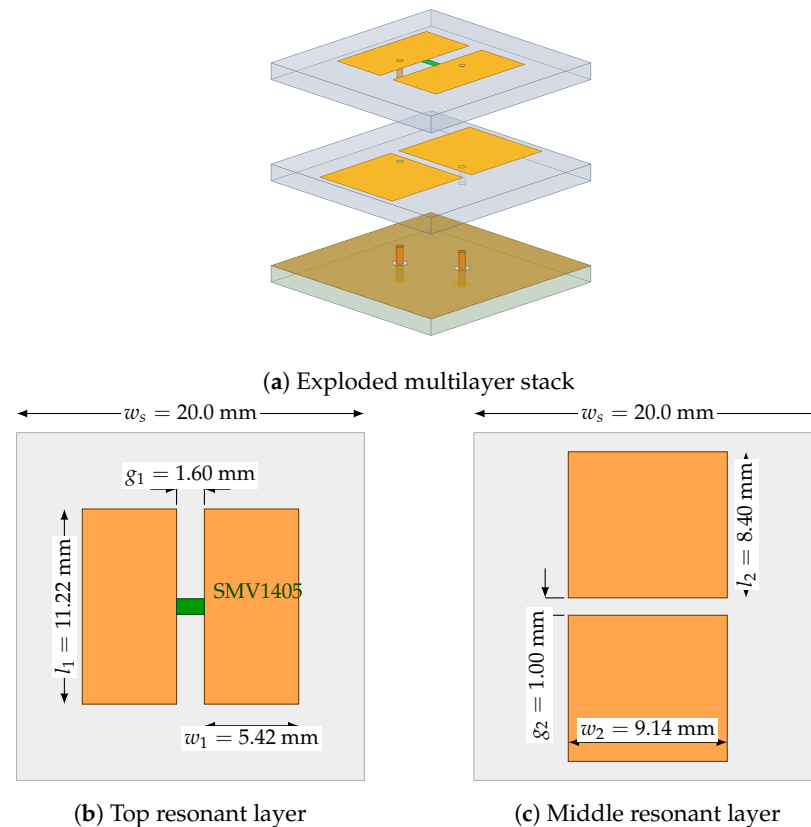


Figure 1. Proposed dual-polarized RIS unit cell: (a) exploded multilayer stack; (b) dimensioned top resonant layer; (c) dimensioned middle resonant layer. Each dielectric layer has thickness $h = 1.57$ mm; the complete parameter set is given in Table 4.

The cell comprises three 1.57-mm-thick dielectric layers. The upper two layers use Rogers RT/duroid 5880, while the bottom FR-4 layer provides mechanical support and carries the varactor bias network. The top patterned copper layer contains a split rectangular patch, and the middle layer contains an orthogonally oriented patch pair. The top face of the bottom layer forms the ground plane, and two plated through-vias connect the upper structure to bias pads on its underside. A Skyworks SMV1405 varactor mounted across the top-patch split is the only active component. Table 4 lists the physical dimensions.

Table 4. Parameter values for the proposed RIS unit cell.

Parameter	Description	Value (mm)
w_s	Unit-cell side (square)	20.0
w_1	Top-layer patch width	5.42
l_1	Top-layer patch length	11.22
g_1	Top-layer split gap	1.6
w_2	Middle-layer patch width	9.14
l_2	Middle-layer patch length	8.4
g_2	Middle-layer gap	1.0
h	Dielectric layer thickness ($\times 3$)	1.57
l_v	Varactor length	2.45
w_v	Varactor width	1.48
d_v	Varactor pad spacing	1.2

The patterned layers and ground plane form a coupled resonant reflector. Changing the varactor reverse-bias voltage changes its capacitance and therefore the complex co- and

cross-polarized reflection coefficients at 5.4 GHz. Because both resonant structures share one varactor bias, each voltage commands one complete polarization response rather than independently selectable TE and TM phases. The patch lengths l_1 and l_2 primarily set the resonant frequencies, while gaps g_1 and g_2 affect the varactor loading.

2.2. Simulation Setup and Electromagnetic Performance

The electromagnetic performance of the proposed unit cell was simulated using Ansys High Frequency Structure Simulator (HFSS; software version not retained; <https://www.ansys.com/products/electronics/ansys-hfss>, accessed on 21 September 2026) with the dimensions listed in Table 4. Figure 2 depicts the model, in which lattice-pair boundaries represent an infinite periodic array. At normal incidence, the simulation included two propagating fundamental modes, corresponding to the x - and y -polarized (0,0) Floquet modes. The resulting S matrix was mapped to the linear-polarization reflection matrix described below.

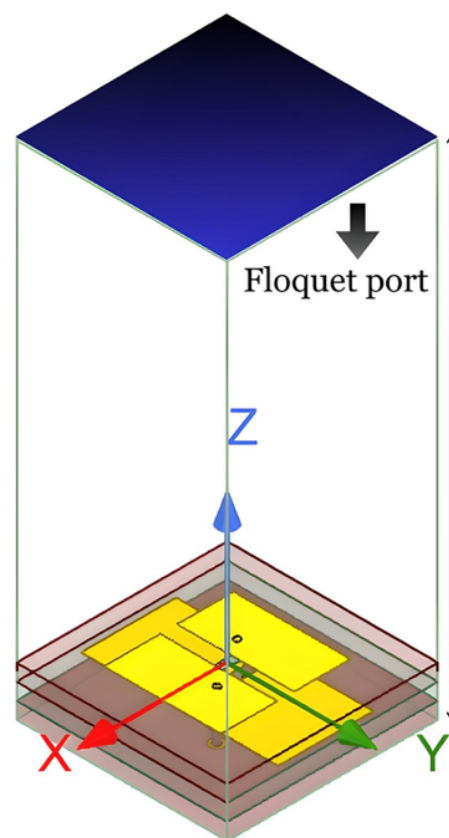


Figure 2. Simulation configuration of the proposed unit cell in Ansys HFSS.

For incident x and y linear polarizations, the simulated reflected field is described by

$$\begin{bmatrix} E_x^r \\ E_y^r \end{bmatrix} = \underbrace{\begin{bmatrix} r_{xx} & r_{xy} \\ r_{yx} & r_{yy} \end{bmatrix}}_{\mathbf{R}} \begin{bmatrix} E_x^i \\ E_y^i \end{bmatrix}. \quad (1)$$

Here, x and y correspond to the TE and TM configurations, respectively. The co-to-cross discrimination is $\text{XPD}_x = 20 \log_{10} |r_{xx}/r_{yx}|$ for x -polarized incidence and $\text{XPD}_y = 20 \log_{10} |r_{yy}/r_{xy}|$ for y -polarized incidence. Positive XPD indicates that the co-polarized reflected component is stronger; the three states evaluated below all exceed 11.9 dB.

Figure 3 and Tables 5 and 6 show that the co-polarized response remains dominant in the corrected model. Across the three representative states, XPD ranges from 12.42 to 13.88 dB for TE incidence and from 11.98 to 13.71 dB for TM incidence. A tighter 2.7-pF solve gives 12.56 and 12.14 dB, respectively, a change below 0.16 dB. The cross-polarized channel contains 3.8–5.0% of incident power or 3.9–6.0% of total reflected power. These simulations support moderate polarization isolation at the evaluated states, but they do not establish independently selectable TE and TM control.

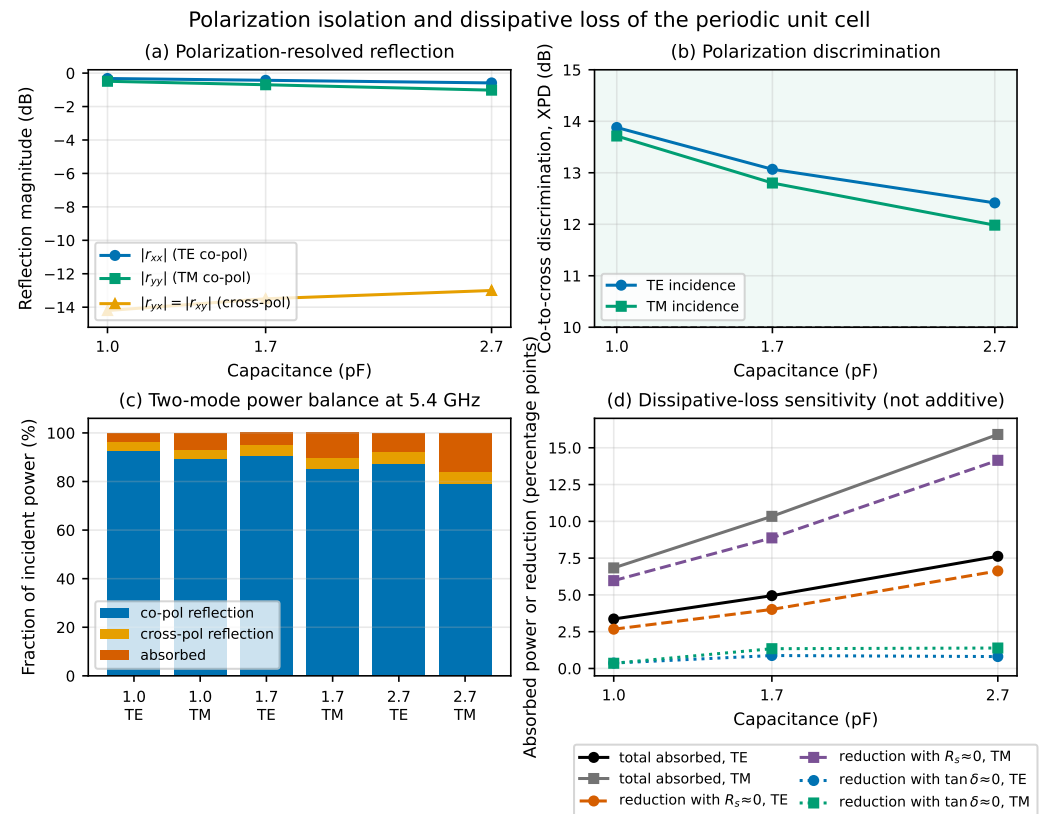


Figure 3. Two-mode periodic-cell results at 5.4 GHz and normal incidence for three representative capacitance states: (a) co- and cross-polarized reflection magnitudes; (b) co-to-cross discrimination; (c) power balance including both reflected polarizations; and (d) one-at-a-time loss-disabled comparisons. The reductions in panel (d) are sensitivities and are not additive component powers.

Table 5. Polarization-resolved reflection and complete two-mode power balance at 5.4 GHz.

C (pF)	Inc.	Co-Pol (dB)	Cross-Pol (dB)	XPD (dB)	Cross-Pol (% Refl.)	P_{abs} (%)	$L_{refl,tot}$ (dB)
1.0	TE	−0.32	−14.20	13.88	3.9	3.36	0.15
1.0	TM	−0.49	−14.20	13.71	4.1	6.84	0.31
1.7	TE	−0.43	−13.50	13.07	4.7	4.95	0.22
1.7	TM	−0.70	−13.50	12.80	5.0	10.34	0.47
2.7	TE	−0.59	−13.00	12.42	5.4	7.62	0.34
2.7	TM	−1.02	−13.00	11.98	6.0	15.91	0.75

The same matrix separates co-polarized attenuation from dissipative loss. For unit incident power, $P_{abs,x} = 1 - |r_{xx}|^2 - |r_{yx}|^2$ and $P_{abs,y} = 1 - |r_{yy}|^2 - |r_{xy}|^2$. When both reflected polarizations are counted, the three tested states return 92.4–96.6% of the incident TE power and 84.1–93.2% of the incident TM power, corresponding to 0.15–0.75 dB total reflection loss. Co-polarized attenuation is 0.32–1.02 dB and exceeds the total reflection loss by only 0.17–0.27 dB because the cross-polarized leakage is modest.

Table 6. One-at-a-time reduction in absorbed power at 5.4 GHz.

C (pF)	Inc.	P_{abs} (%)	$\Delta P_{R_s \rightarrow 0}$ (Points)	$\Delta P_{\tan \delta \rightarrow 0}$ (Points)	Lossless Floor (%)
1.0	TE	3.36	2.67	0.37	<0.01
1.0	TM	6.84	5.97	0.35	<0.01
1.7	TE	4.95	4.01	0.88	<0.01
1.7	TM	10.34	8.88	1.34	0.01
2.7	TE	7.62	6.63	0.81	0.02
2.7	TM	15.91	14.15	1.39	0.04

The baseline model uses $R_s = 0.8 \Omega$ for the varactor, $\tan \delta = 0.0009$ for Rogers 5880, and $\tan \delta = 0.02$ for FR-4. Setting R_s close to zero reduces the absorbed power much more than setting both dielectric loss tangents close to zero, so varactor resistance is the dominant represented dissipative sensitivity. These differences are not an exact additive decomposition because each material change also perturbs the resonant field distribution. All conductors in the recovered model are perfect electric conductors. Consequently, finite-conductivity loss in the patches, vias, and bias network is not quantified; the bias-network geometry is included, but its ohmic loss is not. A tighter 2.7-pF solve changed P_{abs} by less than 0.7 percentage point of incident power.

These two-mode results apply to normal incidence on an infinite periodic cell. Oblique-incidence and finite-array polarimetric behavior remain unquantified. The single-mode supercell study in Section 2.3 is used only to quantify the collective co-polarized response under representative non-uniform loading; it is not used to claim polarization isolation, total dissipative loss, or a polarization-preserving 250° phase span.

To further examine the coupled response, Figure 4 shows the electric-field distribution on the top and middle conducting layers under TE- and TM-oriented excitation.

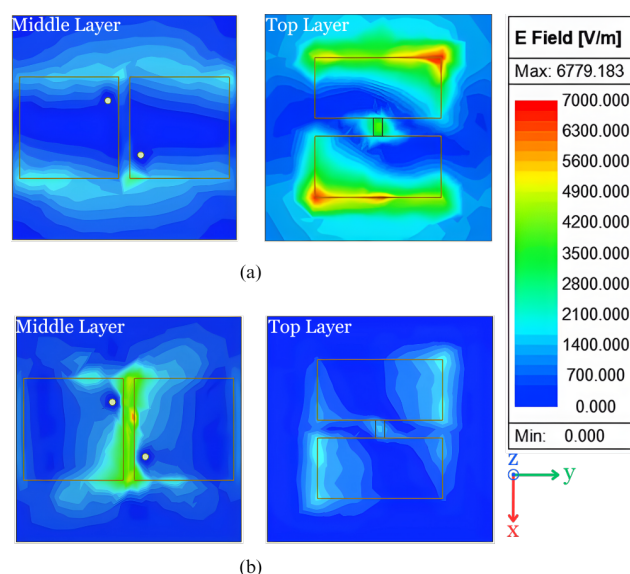
**Figure 4.** Magnitude of the electric-field distribution. (a) middle layer (left) and top layer (right) under TE-oriented excitation. (b) middle layer (left) and top layer (right) under TM-oriented excitation.

Figure 4 provides a qualitative view of the state-dependent field concentration. Under TE (x -polarized) excitation, the field is concentrated on the top-layer split patch containing the varactor. Under TM (y -polarized) excitation, it is concentrated mainly on the middle-layer patches. Residual coupling between the nominally orthogonal responses is quantified by the off-diagonal terms in Equation (1).

2.3. Non-Uniform Neighboring Bias States

A periodically repeated 2×2 supercell was simulated at normal incidence to quantify how unlike neighboring varactor states perturb the collective co-polarized response. Three loading patterns were evaluated for both TE (x -polarized) and TM (y -polarized) incidence: a uniform reference with all cells at 1.8 pF, alternating columns at 1.0 and 2.7 pF (stripes), and a 1.0/2.7-pF checkerboard. The uniform supercell reproduced the corresponding single-cell result to within +0.04 dB and -2.25° for TE and +0.02 dB and -1.13° for TM.

For each 50/50 non-uniform pattern, the coupling-free reference was defined as the complex average $\Gamma_{\text{unc}} = [\Gamma(1.0 \text{ pF}) + \Gamma(2.7 \text{ pF})]/2$. The reported deviations are $\Delta A = 20 \log_{10} |\Gamma_{\text{sc}}/\Gamma_{\text{unc}}|$ and $\Delta\phi = \angle\Gamma_{\text{sc}} - \angle\Gamma_{\text{unc}}$, wrapped to $[-180^\circ, 180^\circ]$, where Γ_{sc} is the simulated supercell reflection coefficient.

Table 7 shows that the two non-uniform TE patterns depart from the uncoupled complex average by 5.82–6.03 dB in magnitude and 58.9 – 68.2° in phase. For TM, the corresponding deviations are 21.16–21.52 dB and 68.0 – 91.6° . The large relative TM magnitude deviations arise because the isolated 1.0- and 2.7-pF coefficients are nearly out of phase, so their uncoupled average is close to a null ($|\Gamma_{\text{unc}}| = 0.0719$); the absolute supercell magnitudes remain 0.822–0.856. These results show that isolated-cell complex averaging can substantially misrepresent the collective response under severe neighboring-state contrasts.

Table 7. Collective co-polarized response of periodically repeated 2×2 non-uniform supercells at 5.4 GHz.

Inc.	Pattern	$ \Gamma_{\text{sc}} $	$\angle\Gamma_{\text{sc}} (^\circ)$	$ \Gamma_{\text{unc}} $	$\angle\Gamma_{\text{unc}} (^\circ)$	$\Delta A \text{ (dB)}$	$\Delta\phi (^\circ)$
TE	Stripes	0.924	-107.02	0.461	-165.94	+6.03	+58.9
TE	Checkerboard	0.902	-97.79	0.461	-165.94	+5.82	+68.2
TM	Stripes	0.856	-35.60	0.0719	-103.62	+21.52	+68.0
TM	Checkerboard	0.822	-12.04	0.0719	-103.62	+21.16	+91.6

Because the 2×2 tile is periodically repeated, the differences in Table 7 are coupling- and supercell-induced deviations rather than pure pairwise mutual-coupling coefficients. The study covers two severe non-uniform patterns and one co-polarized Floquet mode per excitation; it does not quantify the complete neighbor-state space, cross-polarized reflection, or finite-array edge diffraction. Consequently, larger apertures require coupling-aware electromagnetic validation and cell-level or grouped calibration rather than direct extrapolation of isolated periodic-cell states.

3. RIS Fabrication and Control Circuitry

3.1. Waveguide Measurement Comparison

Before full-surface assembly, a fabricated 1×2 sample was examined inside a WR159 rectangular waveguide at nominal capacitance settings of 0.7, 2.0, and 2.7 pF. Figure 5 shows the sample and the measurement setup connected to a vector network analyzer (VNA). At 5.4 GHz, this fixture supports the dominant TE_{10} mode but not the orthogonally polarized TE_{01} mode. The measurements therefore compare the measured and simulated co-polarized responses only at these three discrete settings. They do not validate the complete capacitance-dependent phase trajectory or measure the free-space cross-polarized reflection coefficients in Equation (1).

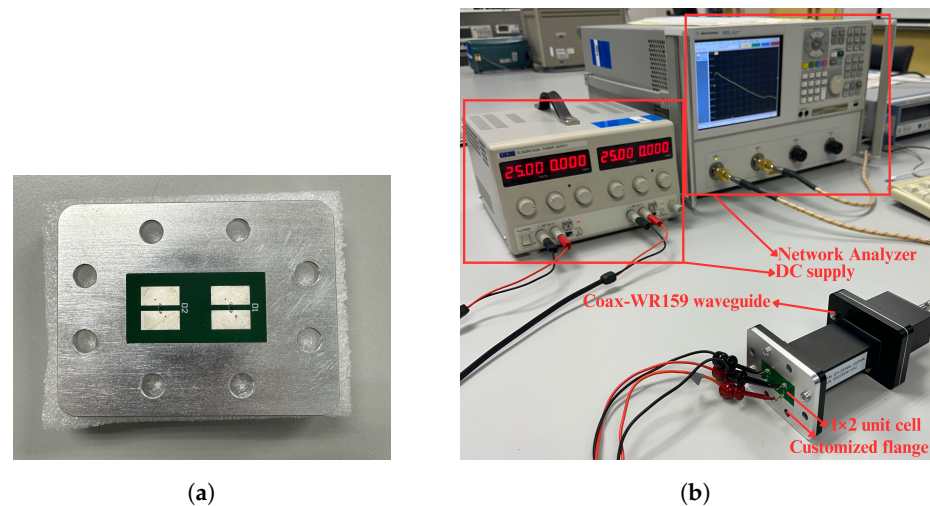


Figure 5. (a) Fabricated 1×2 unit cell sample; (b) Experimental setup for WR159 waveguide measurement.

Figures 6 and 7 compare the simulated and measured waveguide reflection phase and amplitude for the three reported states under TE and TM polarization.

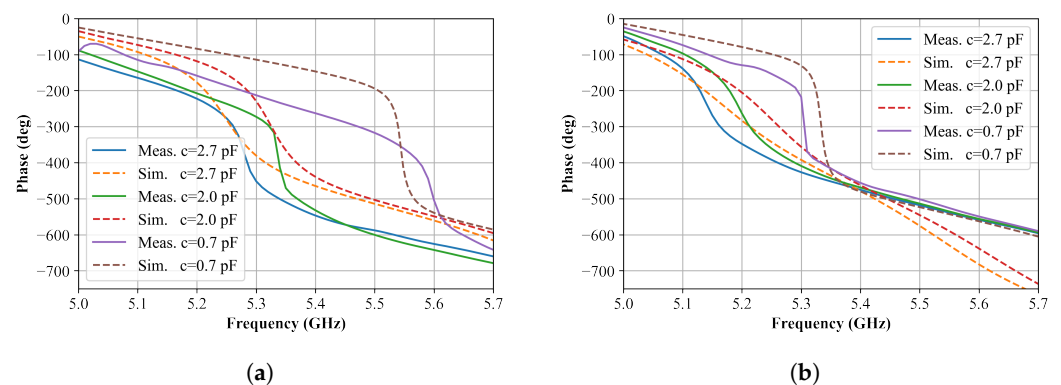


Figure 6. Comparison of simulated and measured reflection phase under: (a) TE polarization; (b) TM polarization.

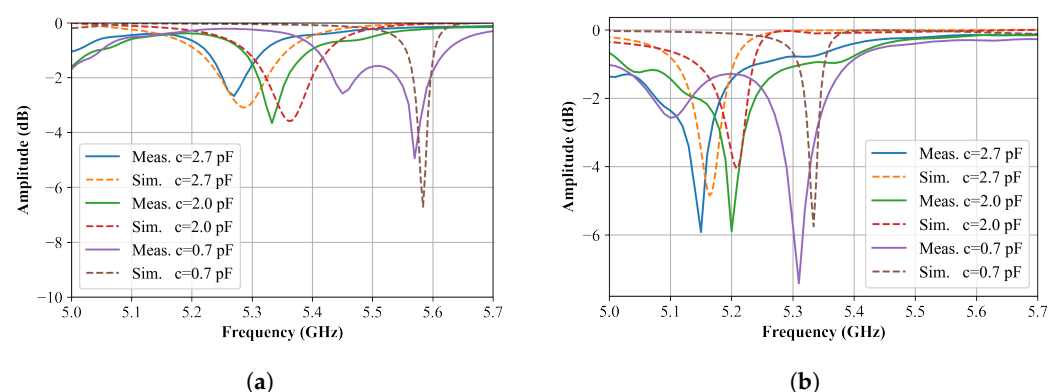


Figure 7. Comparison of simulated and measured reflection amplitude under: (a) TE polarization; (b) TM polarization.

Tables 8 and 9 quantify the comparisons at the three measured capacitance settings. For capacitance settings of 2.7, 2.0, and 0.7 pF, respectively, the circular phase differences at 5.4 GHz are 80° , 95° , and 115° for TE and 5° , 5° , and 25° for TM. The corresponding reflection-amplitude differences are less than 0.1, 1.0, and 0.8 dB for TE and 0.4, 0.5, and 0.8 dB for TM. The resonance-frequency differences are 20, 30, and 10 MHz for TE and 10, 10, and 20 MHz for TM. These comparisons characterize the agreement shown

in Figures 6 and 7 rather than constituting a new uncertainty analysis of the raw VNA measurements. The relatively large TE phase differences and the state-dependent offsets show that the agreement is only partial. Several mechanisms may contribute. The simulation uses periodic boundaries, whereas the measured specimen is a finite 1×2 pair mounted in a waveguide flange. In addition, the WR159 TE₁₀ field is non-uniform across the aperture, unlike the plane-wave excitation of the periodic model. Varactor-model uncertainty, fabrication tolerances, soldering, vias, and finite pad dimensions may also shift the response.

Table 8. Approximate comparison of measured and simulated WR159 reflection phase and amplitude at 5.4 GHz for the TE- and TM-oriented configurations at three nominal capacitance settings.

Pol.	C (pF)	ϕ_{meas} (°)	ϕ_{sim} (°)	$ \Delta\phi $ (°)	A_{meas} (dB)	A_{sim} (dB)	$ \Delta A $ (dB)
TE	2.7	−545	−465	80	−0.4	−0.5	<0.1
TE	2.0	−530	−440	95	−1.0	−2.0	1.0
TE	0.7	−265	−145	115	−0.8	−0.1	0.8
TM	2.7	−475	−480	5	−0.4	0.0	0.4
TM	2.0	−470	−465	5	−0.8	−0.2	0.5
TM	0.7	−455	−480	25	−0.9	0.0	0.8

The phase difference is the magnitude of the circular difference on a 360° interval. Values were read from the displayed curves and rounded to 5° and 0.1 dB; they are not substitutes for a calibrated raw-data uncertainty analysis.

Table 9. Approximate comparison of measured and simulated WR159 resonance frequency for the TE- and TM-oriented configurations at three nominal capacitance settings.

Pol.	C (pF)	$f_{r,\text{meas}}$ (GHz)	$f_{r,\text{sim}}$ (GHz)	$ \Delta f_r $ (MHz)
TE	2.7	5.27	5.29	20
TE	2.0	5.33	5.36	30
TE	0.7	5.57	5.58	10
TM	2.7	5.15	5.16	10
TM	2.0	5.20	5.21	10
TM	0.7	5.31	5.33	20

Values were read from the displayed curves and rounded to 0.01 GHz; they are not substitutes for a calibrated raw-data uncertainty analysis.

3.2. Varactor Diode Characterization and Control Circuitry

Figure 8 shows the typical total-capacitance values tabulated by Skyworks for the SMV1405 SOT-23 single configuration as a function of reverse-bias voltage [30]. A monotone interpolation is used between the tabulated points.

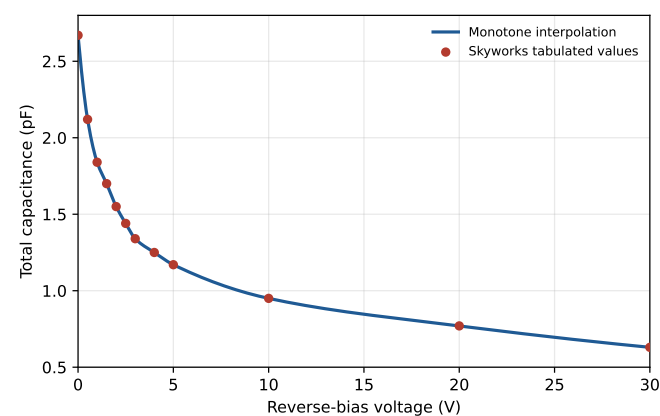


Figure 8. Skyworks SMV1405 tabulated typical capacitance values and the monotone interpolation used for the nominal voltage–capacitance mapping [30].

Each control output is generated by a 12-bit MCP4728 digital-to-analog converter (DAC) followed by an LM324AMX operational-amplifier stage that maps the DAC output to 0–30 V. The 12-bit specification denotes 4096 commanded voltage codes, with an ideal adjacent-code voltage step of $30/4095 = 7.33$ mV; it does not establish 4096 experimentally distinguishable electromagnetic states. Figure 9 shows one four-output circuit section.

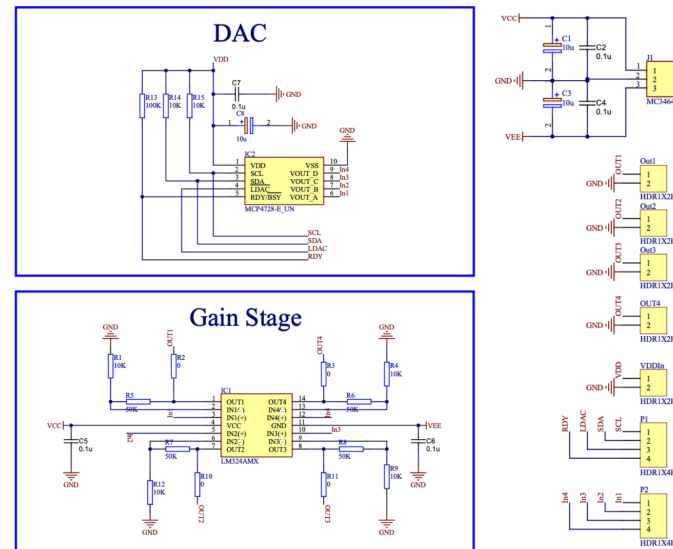


Figure 9. Schematic diagram of the 4-output DAC and operational amplifier control circuit.

An FPGA communicates with the DACs over multiple I²C buses. With the retained host-link setting of 115,200 bps and 400-kbps I²C operation, the reported full-board update interval is approximately 10 ms. This board-update time is distinct from the varactor response time and from the duration of the measurement-based optimization loop. Scaling to more cells would require additional consideration of bus loading, channel count, synchronization, and update scheduling.

The DC consumption of the bias-control electronics and the RF reflection efficiency of the RIS describe distinct power paths. Under maximum-voltage loading, the DACs and operational amplifiers that generate the 36 cell-bias voltages consume 0.21 W and 0.57 W, respectively, for a combined analog-control consumption of 0.78 W. Separately, the full two-mode unit-cell result in Section 2 distinguishes reflected power in both polarization channels from absorbed power. Consequently, low DC control consumption does not by itself establish RF reflection efficiency or end-to-end link efficiency. The latter also depends on aperture size, illumination and spillover, phase and amplitude errors, cross-polarized leakage, propagation geometry, and the direct Tx–Rx path.

Table 10 gives a first-order scaling estimate for independently controlled cells. Because both the MCP4728 and LM324AMX provide four channels per package, an N -cell array requires $\lceil N/4 \rceil$ packages of each type and N bias lines. The power estimates scale the measured 36-channel DAC and amplifier values linearly with channel count.

Table 10. First-order scaling of the analog bias-control hardware. Power values include only the DAC and amplifier stages.

Cells	DAC Packages	Amplifier Packages	Bias Lines	Analog Power (W)
36	9	9	36	0.78
100	25	25	100	2.17
256	64	64	256	5.55
1024	256	256	1024	22.19

These estimates exclude the FPGA, regulators, power-conversion losses, and communication interfaces and therefore are not total controller-power predictions. With a fixed number of serial buses, configuration-transfer time grows approximately with the number of channels; additional parallel buses or distributed controllers can reduce this scaling. Larger arrays would consequently require partitioned bias electronics, synchronization across controller sections, power distribution, and thermal design. In particular, the first-order 1024-cell analog estimate indicates that heat dissipation could no longer be treated as negligible. Exact FPGA-resource scaling depends on the selected bus architecture and is not inferred from the prototype.

3.3. Mechanical CAD Design and Structural Integration

The 6×6 RIS array was constructed using a modular holder structure designed in computer-aided design (CAD) software and 3D-printed. Figure 10 identifies the complete assembly, supporting base, and unit-cell mounting frame.

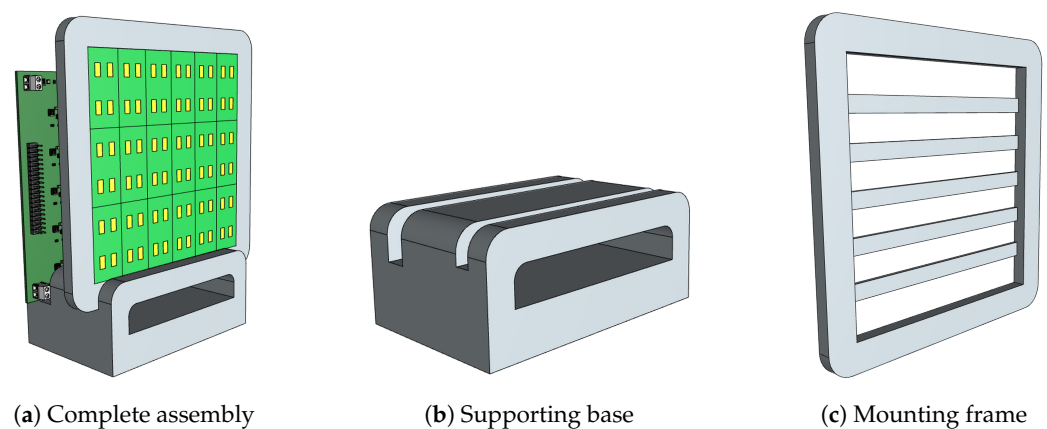


Figure 10. Mechanical CAD views of the RIS prototype: (a) complete assembly with the 6×6 panel and control PCB; (b) supporting base and board guides; (c) unit-cell mounting frame. Component names are typeset as vector subcaptions for legibility at the final journal size.

The 6×6 arrangement of 20-mm cells forms a $120 \text{ mm} \times 120 \text{ mm}$ aperture, approximately 2.2λ per side at 5.4 GHz. This prototype size kept the channel count at 36 and accommodated the modular 1×2 fabrication. The experiments evaluate changes at fixed receiver locations; no angular radiation pattern was measured. Larger apertures would require more bias channels and a separate analysis of coupling, control scaling, and thermal requirements.

Fabrication, assembly, and varactor tolerances can shift the voltage-dependent response of individual cells, as suggested by the state-dependent offsets in the waveguide comparison. For a larger implementation, the nominal voltage mapping should therefore be treated as an initial lookup rather than an exact cell-independent calibration. A practical approach is to characterize each cell, or groups of statistically similar cells, over the operating-voltage range and store the resulting voltage-to-response lookup data in the controller. Closed-loop link optimization can then refine those settings in the installed propagation environment.

4. Experimental Performance Optimization

4.1. Transmission Power Maximization

The fabricated 6×6 RIS was evaluated in a free-space setup using two horn antennas connected to a VNA for fixed-receiver S_{21} optimization. The retained experimental description places each horn aperture at a nominal radial spacing of 30 cm from the RIS, with

both propagation directions 30° from the RIS surface normal. Figure 11 pairs the original photograph with a plan-view schematic that defines these distances and angles.

At 5.4 GHz, the free-space wavelength is approximately 55.5 mm. For the 120-mm aperture, the conventional Fraunhofer distance $R_{FF} = 2D^2/\lambda$ is approximately 0.52 m. The nominal 0.30-m antenna–RIS spacing in this experiment is therefore inside this boundary and is treated as radiating near-field operation.

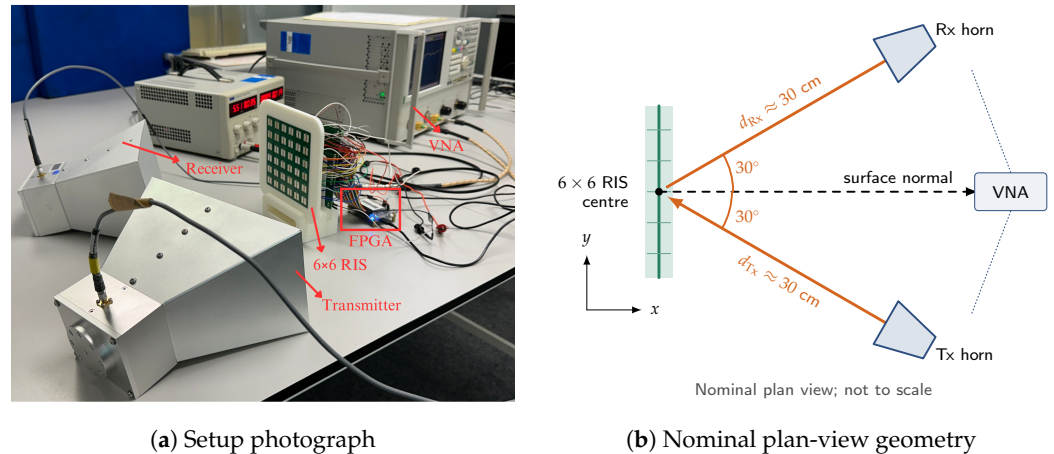


Figure 11. Fixed-receiver transmission-coefficient ($|S_{21}|$) experiment: (a) original VNA-based setup photograph; (b) nominal plan view showing the Tx and Rx paths, the two 30 cm antenna–RIS spacings, and the two 30° angles relative to the RIS surface normal.

A standard implementation of the established non-dominated sorting genetic algorithm II (NSGA-II) [31] was used as a population-based search tool for increased and decreased $|S_{21}|$ at the fixed receiver. The optimization algorithm is not proposed as a contribution of this work. Each candidate contained 36 commanded cell-bias values, one for each RIS cell. The retained experimental description specifies a population of 50 and a fixed stopping budget of 10 generations for the $|S_{21}|$ searches. Figure 12 shows the recorded TE and TM curves for the reference and optimized configurations.

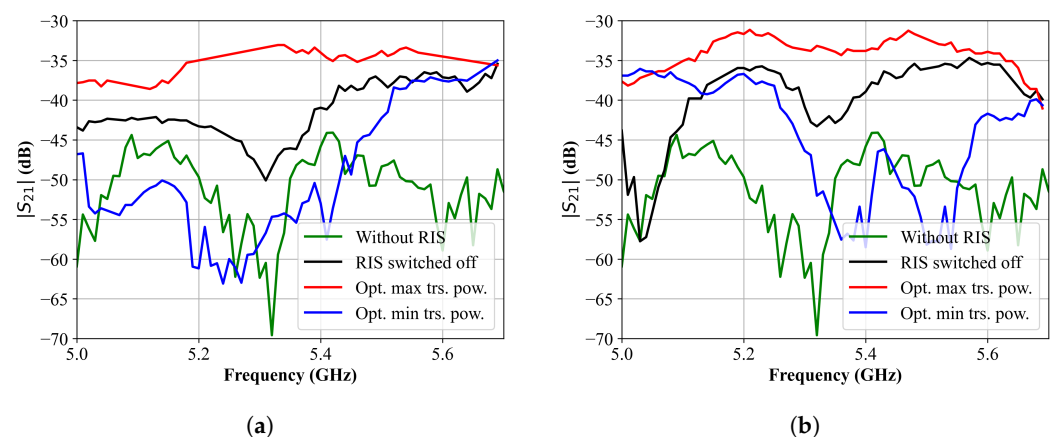


Figure 12. Comparison of $|S_{21}|$ at the fixed receiver for (a) TE and (b) TM configurations. “Without RIS” denotes removal of the surface; “RIS switched off” denotes the surface remaining in place with its control electronics unpowered and no commanded reverse-bias voltages applied. “Opt. max/min trs. pow.” denotes the recorded configurations obtained by searches for increased/decreased transmission.

The two reference conditions are physically distinct. For “Without RIS,” the surface was removed from the setup. For “RIS switched off,” the surface remained in position while its control electronics were unpowered and no commanded reverse-bias voltages were applied; the unpowered structure could therefore still interact passively with the

incident field. At 5.4 GHz, the recorded maximizing and minimizing configurations produce different $|S_{21}|$ values at the fixed receiver. This contrast demonstrates received-power redistribution at that location, but it does not constitute an angular beam-steering measurement. Because the link also contains a direct Tx–Rx contribution, the observed change cannot be interpreted as the gain of the RIS path alone.

The purpose of this experiment was to demonstrate near-field received-power optimization at a fixed receiver, rather than to characterize far-field radiation-pattern quantities such as sidelobe level, beam-pointing error, or pattern-based beamforming efficiency. Because the experiment optimized the measured fixed-receiver $|S_{21}|$ directly, no ideal scalar phase-truncation model was used in the search. A quantitative near-field array study incorporating the polarization-dependent response of every commanded state remains future work.

4.2. SNR Optimization in Anechoic Chamber

To evaluate real-time end-to-end communication performance, an anechoic-chamber testbed was assembled as shown in Figure 13. The retained annotated photograph records nominal Tx–RIS and RIS–Rx spacings of 84 cm, with the RIS at the vertex of the 75° included angle between the two propagation directions. Figure 13b makes this angular definition explicit.

The nominal 0.84-m antenna–RIS spacing is greater than the approximately 0.52-m Fraunhofer distance calculated above and is therefore beyond the conventional far-field boundary of the prototype aperture.

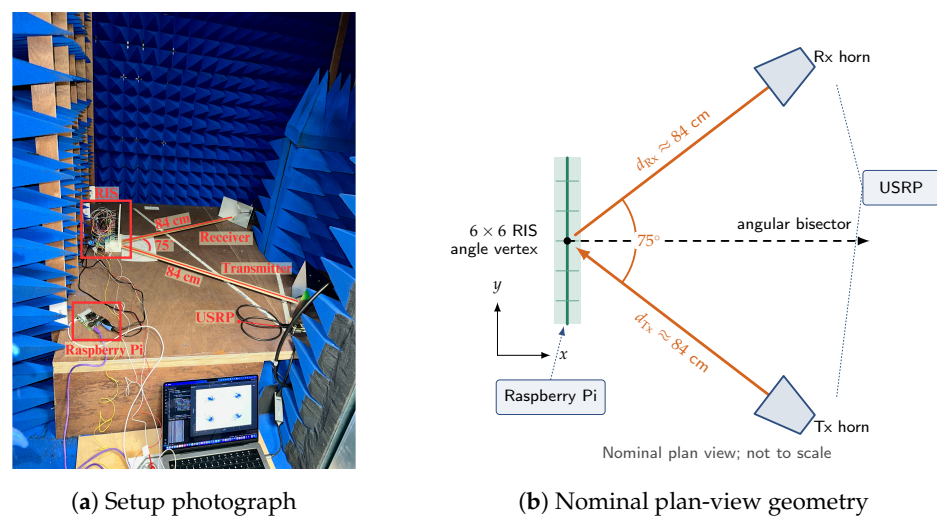


Figure 13. Anechoic-chamber setup for closed-loop SNR optimization: (a) original setup photograph; (b) nominal plan view showing the Tx and Rx paths, the two 84 cm antenna–RIS spacings, and the 75° included angle with the RIS at its vertex.

The signal processing architecture using a Universal Software Radio Peripheral (USRP B205mini; Ettus Research, Santa Clara, CA, USA) and Raspberry Pi controller running GNU Radio (software version not retained; <https://www.gnuradio.org/>, accessed on 21 September 2026) is illustrated in Figure 14.

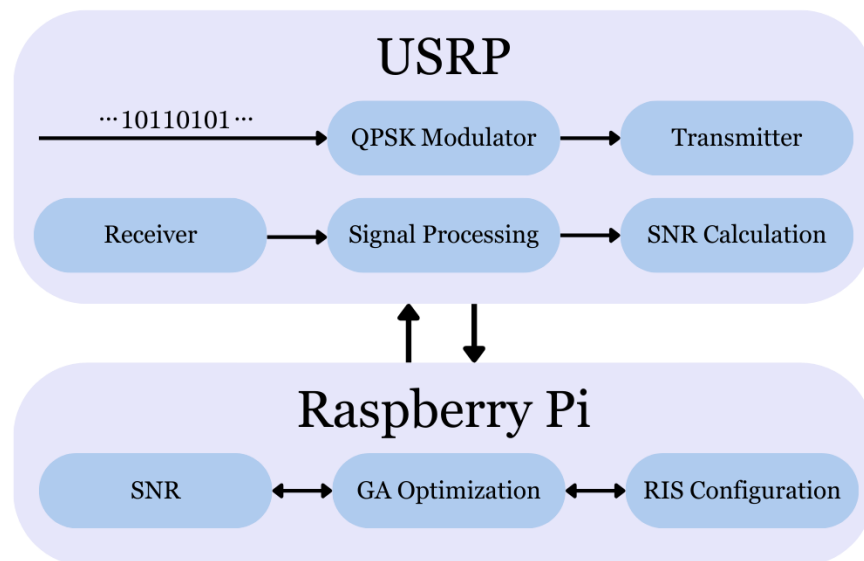


Figure 14. SNR optimization process using USRP for signal processing and Raspberry Pi for real-time RIS configuration.

The same standard NSGA-II method was used as the search engine in the closed-loop SNR experiment, with received SNR as the measured objective. We cite the established algorithm instead of reproducing its derivation or pseudocode because no modification to NSGA-II is proposed. The surviving records do not establish the exact initialization distribution, parent-selection setting, crossover and mutation operators or probabilities, elitism setting, or random seed used in either experiment. These source-level settings are therefore not reconstructed.

Figure 15 shows the single retained SNR-optimization trajectory for each polarization configuration at 5.4 GHz.

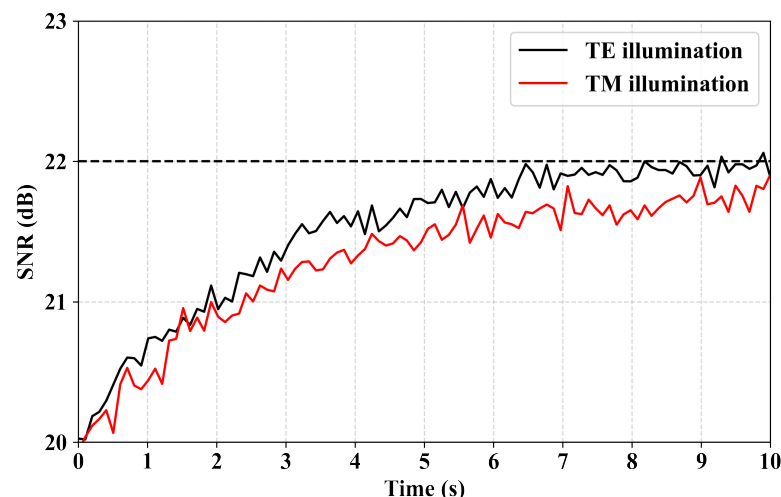


Figure 15. Single retained SNR-optimization trajectory for each polarization configuration at 5.4 GHz. The curves are referenced to the first recorded optimization state, not to a separately measured RIS-off or RIS-absent condition.

In these recorded trajectories, the SNR rises from approximately 20 dB to nearly 22 dB for TE and 21.7 dB for TM within about 6 s. The approximately 2 dB result is therefore a within-run change from the first recorded optimization state to a later optimized state. The exact initial 36-cell bias vector was not retained, and no separate RIS-off or RIS-absent SNR trace is available.

For the present experiment, let $q \in \{\text{initial, optimized}\}$ identify the two recorded RIS states. The appropriate coherent model is $E_{\text{rx}}^{(q)} = E_{\text{d}} + \sum_{n=1}^{36} E_n^{(q)}$, where E_{d} is the direct field and a spherical-wave cell contribution can be written as $E_n^{(q)} = K_n \Gamma_n^{(q)} \exp[-jk(r_{\text{T},n} + r_{\text{n},\text{R}})] / (r_{\text{T},n} r_{\text{n},\text{R}})$. Here, K_n contains the antenna-pattern, element-pattern, and polarization factors, while $\Gamma_n^{(q)}$ is the state-dependent reflection coefficient. The observable change is consequently $\Delta P = 20 \log_{10} |E_{\text{rx}}^{(\text{optimized})} / E_{\text{rx}}^{(\text{initial})}|$, not a function of element count alone.

5. Conclusions

This paper presented a 5.4-GHz single-varactor, dual-polarized RIS with shared-bias analog tuning. A two-mode Floquet analysis of three representative capacitance states gives 11.98–13.88 dB co-to-cross discrimination, with only 3.8–5.0% of incident power returned in the cross-polarized channel. The result supports moderate polarization isolation at normal incidence while preserving the important qualification that a single shared bias does not provide independently selectable TE and TM states. A separate WR159 comparison used nominal capacitance settings of 0.7, 2.0, and 2.7 pF and only the supported TE₁₀ waveguide mode.

The assembled 6×6 surface was examined in two proof-of-concept, fixed-receiver experiments. The recorded $|S_{21}|$ curves show received-power redistribution between maximizing and minimizing configurations. The single retained SNR trajectory for each polarization configuration shows an approximately 2 dB within-run rise from its first recorded state. The DAC and amplifier stages used to generate the 36 bias voltages consumed a combined 0.78 W under maximum-voltage loading.

The two-mode power balance was evaluated for both TE and TM incidence at varactor capacitances of 1.0, 1.7, and 2.7 pF. Across these six simulated polarization–capacitance cases, total reflected power is 84.1–96.6% of the incident power, absorbed power is 3.4–15.9%, and co-polarized attenuation is 0.32–1.02 dB. The power balance therefore distinguishes attenuation caused by absorption from power converted to the cross-polarized channel. One-at-a-time comparisons identify varactor series resistance as the dominant dissipative sensitivity represented in the model. A separate periodically repeated 2×2 study showed substantial collective-response deviations under severe stripe and checker-board loading: 5.82–6.03 dB and 58.9–68.2° for TE and 21.16–21.52 dB and 68.0–91.6° for TM relative to the uncoupled complex average. The TM magnitude ratio is amplified by the near-null uncoupled reference and is not an array-gain measure. The architecture therefore combines one active component per cell and low analog-control consumption with moderate simulated polarization isolation under the evaluated conditions. Dense two-mode state sweeps, oblique-incidence, finite-array polarimetry, tolerance analysis, and measured polarimetric validation remain future work.

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