

A Compact Wideband Slotted Annular-Ring MIMO Antenna with Front-Side Parasitic Decoupling and Circuit-Based Analysis for X-/Ku-Band Applications

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ABSTRACT: This paper reports a two-port multiple-input multiple-output (MIMO) antenna that pairs a slotted annular-ring radiator with a single front-side parasitic strip for X- and Ku-band applications. Three features distinguish the design. A set of eight arc slots elongates the ring-current path by 9.1% and drags the lower band edge down to 8.00 GHz on a board only $0.40\lambda_o$ tall, so the aperture keeps a narrow-strip outline suited to edge mounting. The front-side strip works as a three-quarter-wave trap on the shared substrate and lifts worst-case in-band isolation by 8.4 dB with no defected ground and no via. Every element of the accompanying circuit model maps to a named piece of metal, which lets the model track S_{21} as well as S_{11} , whereas most published models fit the reflection coefficient alone. The radiator occupies $40 \times 15 \times 0.8 \text{ mm}^3$ on Rogers RT/Duroid 5880 ($\epsilon_r = 2.2$, $\tan \delta = 0.0009$). Measurement gives a -10 dB band of 8.00–13.96 GHz, a fractional bandwidth of 54.3%, isolation better than 21.1 dB across the band with a 33.4 dB peak, gain of 4.0–6.5 dBi, and radiation efficiency of 77–84%. Diversity metrics stay within useful limits: envelope correlation coefficient below 0.0038, diversity gain above 9.981 dB, mean effective gain between -3.05 and -3.52 dB , channel capacity loss under 0.331 b/s/Hz and ergodic capacity of 13.01–13.28 b/s/Hz. Characteristic mode analysis at 8.85 and 11 GHz identifies modes carried by the ring. A quarter-wave counterpoise argument predicts the lower edge to within 1.11% of measurement, and two independent routes to the ring resonance differ by only 4.14%. The circuit model reproduces the full-wave response with 0.352 dB RMS error on S_{11} and 0.611 dB on S_{21} , matching the upper band edge to 0.07%.

1. INTRODUCTION

X- and Ku-band links carry satellite downlinks, marine and airborne radar, and point-to-point backhaul. Terminals for these services increasingly need spatial diversity, which pushes designers toward multi-port apertures on a single laminate. Two problems then arise together. The elements must share a small board, so mutual coupling rises; the band of interest spans roughly 8 to 14 GHz, so any decoupling structure must work across an octave rather than at a spot frequency.

Most reported remedies attack the coupling path through the ground. Defected ground structures, electromagnetic band-gap patterns, and slotted counterpoises all interrupt surface current and perturb the radiating aperture at the same time. The cost is usually paid in the back lobe or in cross-polarisation. Parasitic strips placed on the radiating layer avoid that trade because they do not cut the counterpoise — the difficulty is that a plain strip resonates at one frequency and stops helping a few hundred megahertz away.

A second gap concerns how these antennas are explained. Characteristic mode analysis (CMA) has become a standard tool for choosing where to feed a multi-port radiator, and lumped circuit models are widely fitted to the reflection coefficient afterwards. The two are rarely joined, and the circuit models that do appear are almost always one-port. A one-port model provides no information about isolation, which is the quantity a MIMO designer cares about most.

This paper addresses both gaps in one structure. The radiator is an annular ring cut with two concentric rings of four arc slots, fed by a high-impedance microstrip line over a truncated ground. A single strip on the radiating layer sits between the two elements. The contributions are as follows.

1. A slotted annular-ring element is synthesised from closed-form expressions rather than from parameter sweeps, and the lower band edge predicted by a quarter-wave counterpoise argument lands within 1.11% of the measured value.
2. Two independent analytical routes to the ring resonance — a mean-perimeter condition and a circular-patch condi-

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tion corrected for fringing — are shown to differ by only 4.14%, which turns the arc-slot elongation claim into a quantified mechanism rather than an assertion.

3. The front-side parasitic strip is identified as a three-quarter-wave trap on a two-section substrate, its predicted resonance falling 4.3% from the measured isolation peak, and it raises worst-case in-band isolation by 8.4 dB without touching the ground plane.
4. A two-port circuit model is built by walking the physical signal path, with the feed line and exposed neck fixed from closed-form expressions and only the radiating and coupling branches fitted. The model reproduces S_{11} and S_{21} together.
5. CMA at two in-band frequencies links the measured behaviour to specific modal currents and their radiation patterns.

2. RELATED WORK AND RESEARCH GAP

2.1. Wideband Multi-Port Apertures for X- and Ku-Band

Singh et al. reported a two-port mushroom-shaped patch pair of $27 \times 23 \times 1.6 \text{ mm}^3$ that produces several narrow X-band responses from one port and a 13.5–15.8 GHz wideband response from the other, decoupled by a defected ground [1]. A quad-port design covering 8.5–18.5 GHz used an inverted H-shaped parasitic element shared between radiators to hold isolation above 15 dB [2]. Mistri et al. obtained three wide bands beginning at 10.2–18.4 GHz from a dual-port snake-head patch on a partial ground, with a microstrip link between the elements supplying decoupling [3]. Yadav et al. placed a tilted isolation stub at 45° behind a sickle-shaped pair and reached a 19.4% band at 13.07–15.87 GHz within a quad-band response [4]. Saleem et al. set two slotted circular radiators orthogonally on RT/Duroid 5880 and reported 3–13 GHz operation with better than 25 dB isolation and 6.5 dB peak gain at 8.5 GHz [5]. A quad-port millimetre-wave array on the same laminate delivered 24–40 GHz with isolation above 44 dB using diamond slots and a defected ground geometry [6].

2.2. Decoupling without Disturbing the Counterpoise

Parasitic and stub-based decoupling avoids cutting the ground. Xi et al. combined a defected ground with a slotted stub to redirect the ground current in a dual-band two-port array [7]. A T-stub with a modified ground served a two-port multi-standard design across several sub-6 GHz allocations [8]. Sim et al. inserted a dollar-shaped parasitic structure between dual-band elements [9]. Qin et al. suppressed coupling in a compact wideband pair for 5G MIMO without an added decoupler [10]. Fang et al. placed a decoupling structure between stacked patches [11] and later demonstrated a self-decoupling shared-aperture array of eight elements [12]. Kulkarni et al. reduced mutual coupling in a dual circularly polarised quad-port X-band array [13]. A parasitic T and inverted-L strip pair produced dual stopbands and improved isolation by more than 31 dB in

a quad-element geometry [14]. Prasad and Makkapati compared slotted-ground and M-shaped defected ground cells for the same purpose [15].

2.3. Modal and Circuit-Level Explanation

Thackston et al. built a synthesis loop that selects feed positions from modal significance curves computed in a CMA solver [16]. Khan et al. used CMA to raise both gain and isolation in a quad-element sub-6 GHz array [17]. CMA has also guided a circularly polarised ultra-wideband monopole [18] and the excitation of orthogonal modes on a single substrate in the millimetre-wave band [19]. On the circuit side, Meher and Mishra derived a mathematical equivalent circuit for a broadband circularly polarised semi-annular ring monopole [20], and lumped models have been fitted to concentric semi-circular ring radiators [21] and to circular-ring monopoles with tapered feeds [22].

2.4. Research Gap

Three gaps follow from the survey. First, decoupling for octave-wide X- and Ku-band pairs still relies mostly on ground modification [1, 3, 15], which perturbs the aperture; the parasitic-strip alternative [9–12] has been demonstrated mainly on narrower sub-6 GHz bands, and no reported X/Ku pair explains its strip as a resonator with a predicted resonance. Second, arc-slot loading used to lower the resonance of ring radiators is normally presented as a swept parameter, so the size reduction it buys is never separated from the counterpoise effect. Third, circuit models for MIMO antennas remain one-port [20–22]; none of the works surveyed reproduces the measured isolation from a lumped model, and none couples such a model to a modal analysis of the same structure.

The proposed antenna addresses each gap in turn. The decoupling element is a strip on the radiating layer, and its three-quarter-wave resonance on a two-section substrate is computed in closed form and checked against the measured isolation peak. The arc-slot elongation is quantified as a 9.1% path increase by comparing the analytical ring resonance with the measured lower edge, so the slot contribution and counterpoise contribution are reported separately. The circuit model is a two-port built by physical partition, and it reproduces the measured isolation notch, as well as the impedance band; CMA at two in-band frequencies provides the modal picture that the circuit branches represent.

3. ANTENNA GEOMETRY AND CLOSED-FORM DESIGN

Figure 1 gives the geometry. Two identical elements sit on Rogers RT/Duroid 5880 ($\epsilon_r = 2.2$, $\tan \delta = 0.0009$) of thickness $h = 0.8 \text{ mm}$, on a board of $W = 40 \text{ mm}$ by $L = 15 \text{ mm}$. Each element is an annular ring of outer radius $R_1 = 5.6 \text{ mm}$ with a central circular aperture of radius $R_2 = 2.45 \text{ mm}$. Two concentric rings of four arc slots are cut into the annulus, dimensioned by L_4 to L_8 and by the arc radius R_3 . A microstrip line of width $W_1 = 1.5 \text{ mm}$ and length $L_1 = 3.45 \text{ mm}$ feeds the ring from the lower edge. The reverse side carries two par-

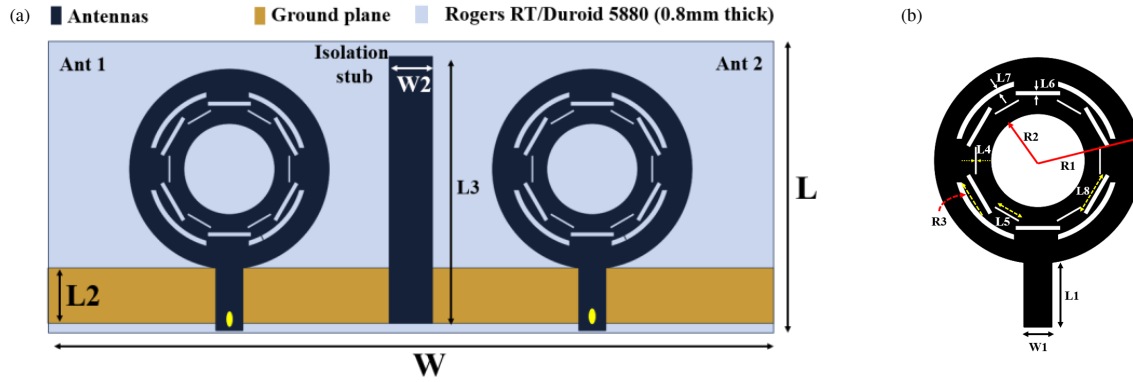


FIGURE 1. (a) Overall structure and dimensions of the proposed antenna and (b) detailed dimensions of a single antenna element.

TABLE 1. Various design parameters of the proposed antenna.

Parameters	L	L ₁	L ₂	L ₃	L ₄	L ₅
Values (mm)	15	3.45	3.05	14.7	0.13	1.5
Parameters	L ₆	L ₇	L ₈	W	W ₁	W ₂
Values (mm)	0.25	0.317	2.4	40	1.5	2.4
Parameters	R ₁	R ₂	R ₃			
Values (mm)	5.6	2.45	1.75			

tial ground planes of height $L_2 = 3.05$ mm. A parasitic strip of width $W_2 = 2.4$ mm and length $L_3 = 14.7$ mm occupies the centre of the radiating layer between the elements. Table 1 lists all values.

The stack sets two effective permittivities. Metal above the truncated ground sees the microstrip value; metal beyond the ground edge sees the quasi-static average of substrate and air,

$$\varepsilon_{eff,0} = \frac{\varepsilon_r + 1}{2} = 1.600 \quad (1)$$

$$\varepsilon_{eff,m} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left(1 + \frac{12h}{W_1} \right)^{-1/2} = 1.8206 \quad (2)$$

The feed impedance follows from the Hammerstad expression for $W_1/h \geq 1$,

$$Z_{0f} = \frac{120\pi}{\sqrt{\varepsilon_{eff,m}} \left[\frac{W_1}{h} + 1.393 + 0.667 \ln \left(\frac{W_1}{h} + 1.444 \right) \right]} \quad (3)$$

which returns 68.68Ω . A 50Ω line on this laminate would need $W_1 = 2.465$ mm. The narrower line is deliberate: at 3.05 mm it is $0.153\lambda_g$ long at band centre, so it acts as a short transformer section between the 50Ω reference and the ring admittance rather than as a matched feed. That transformation is one reason that the -10 dB band is wide.

3.1. Lower Band Edge

Beyond the ground edge, the ring and its counterpoise form an asymmetric dipole. The end-to-end conducting length is the feed length plus the ring diameter, and the first resonance is a half-wave condition on that length,

$$f_L = \frac{c}{2(L_1 + 2R_1)\sqrt{\varepsilon_{eff,0}}} \quad (4)$$

With $L_1 + 2R_1 = 14.65$ mm, Eq. (4) gives a lower band edge of 8.089 GHz, which is 1.11% from the measured lower edge of 8.00 GHz and 5.87% from the simulated 8.593 GHz.

3.2. Ring Resonance by Two Independent Routes

The mid-band resonance is a property of the ring itself, and it can be reached in two ways. Treating the metal as a circular patch and correcting the radius for fringing,

$$a_e = R_1 \left[1 + \frac{2h}{\pi R_1 \varepsilon_r} \left(\ln \frac{\pi R_1}{2h} + 1.7726 \right) \right]^{1/2} \quad (5)$$

$$f_{TM_{11}} = \frac{1.8412c}{2\pi a_e \sqrt{\varepsilon_r}} \quad (6)$$

gives $a_e = 6.0635$ mm and 9.768 GHz. Treating it instead as an annular ring resonating when the mean circumference equals one guided wavelength,

$$f_{ring} = \frac{c}{\pi(R_1 + R_2)\sqrt{\varepsilon_{eff,0}}} \quad (7)$$

gives 9.372 GHz. The two routes differ by 4.14% , and their mean of 9.570 GHz sits 8.1% above the lower CMA frequency of 8.85 GHz. Agreement between two arguments that share no assumptions is what makes the next step defensible.

3.3. Arc-Slot Path Elongation

The arc slots cut the annulus into narrow concentric tracks, so the surface current must detour around each slot end. The elongation follows from the ratio of the unperturbed ring resonance to the realised lower edge,

$$k = \frac{f_{ring}}{f_{L,sim}} = \frac{9.372}{8.593} = 1.0906 \quad (8)$$

so the slots lengthen the effective path by 9.1% . The benefit is a lower band edge for the same outer radius — the trade is a narrower conducting track, which raises the branch impedance of the radiator and shows up directly in the circuit model of Section 6.

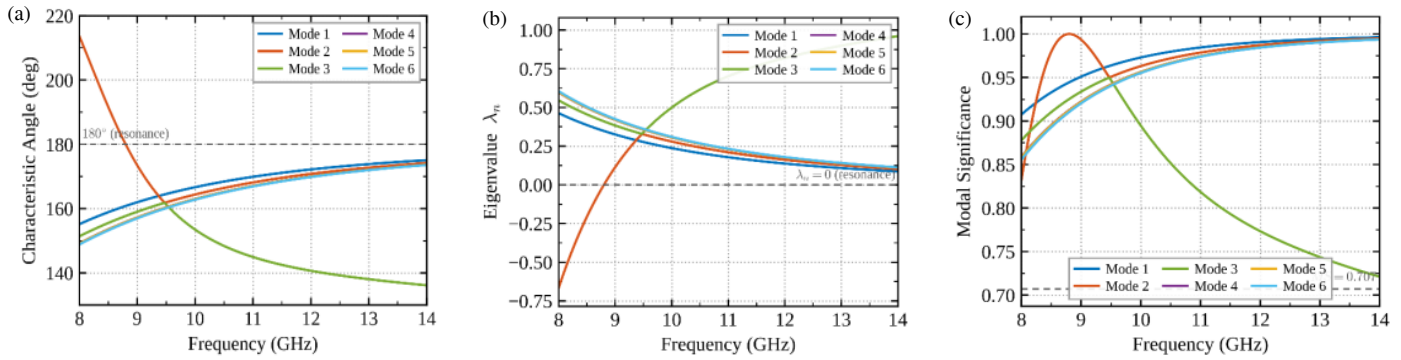


FIGURE 2. Characteristic mode analysis at 8.85 GHz. (a) Characteristic angle, (b) eigenvalue, and (c) modal significance.

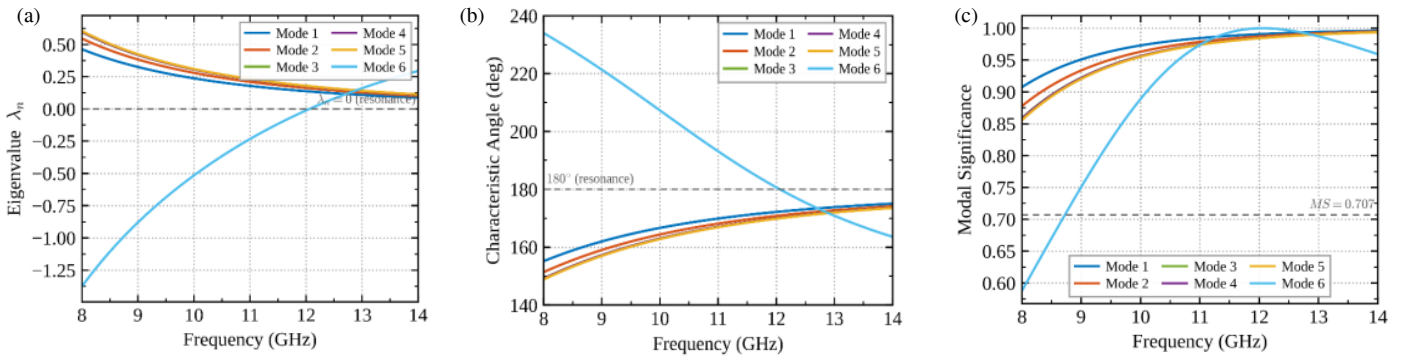


FIGURE 3. Characteristic mode analysis at 11 GHz. (a) Characteristic angle, (b) eigenvalue, and (c) modal significance.

3.4. Upper Band Edge and the Counterpoise

The upper edge is governed by the truncated ground, not by the ring. Current turning around the board edge sees the ground height augmented by the substrate thickness, and a quarter-wave condition on that length gives

$$f_H = \frac{c}{4(L_2 + h) \sqrt{\epsilon_{eff,m}}} \quad (9)$$

For $L_2 = 3.05$ mm, Eq. (9) predicts 14.43 GHz against a simulated 13.671 GHz, an error of 5.6%. Equation (9) likewise predicts 30.03, 19.49, and 11.45 GHz for L_2 of 1.05, 2.05, and 4.05 mm, and Section 4 shows that the swept response follows those predictions.

3.5. Parasitic Strip Resonance

The strip runs over ground for $L_2 = 3.05$ mm and over bare substrate for $L_3 - L_2 = 11.65$ mm, so its electrical length accumulates at two rates,

$$\theta(f) = \frac{2\pi f}{c} (\sqrt{\epsilon_{eff,m}} L_2 + \sqrt{\epsilon_{eff,0}} (L_3 - L_2)) \quad (10)$$

Setting $\theta = \pi$, $3\pi/2$, and 2π gives 7.951, 11.927, and 15.903 GHz. The measured isolation peak falls at 11.43 GHz, which is 4.3% from the three-quarter-wave value. The strip therefore works in band as a $3\lambda/4$ trap, not as the half-wave resonator that a length-only estimate would suggest.

3.6. Characteristic Mode Analysis

CMA separates the metal's radiating behaviour from the feed, so it shows which currents the ring can support before any port is attached. The eigenvalue λ_n , characteristic angle α_n and modal significance MS_n are related by [23]

$$\alpha_n = 180^\circ - \tan^{-1}(\lambda_n), \quad MS_n = \left| \frac{1}{1 + j\lambda_n} \right| \quad (11)$$

A mode resonates where $\lambda_n = 0$, equivalently where $\alpha_n = 180^\circ$ and modal significance peaks at unity.

Figures 2 and 3 show the characteristic angle, eigenvalue, and modal significance at 8.85 and 11 GHz and the two frequencies that bracket the analytical ring resonance in Subsection 3.2. Several modes reach resonance ($\lambda_n = 0$, $\alpha_n = 180^\circ$) in the band, which is consistent with the wide match: a single mode cannot support a 45% band at this Q. The modal significance curves separate cleanly, so the modes the feed excites can be distinguished from those it does not.

Figures 4 and 5 give the three-dimensional modal far fields. The lower-order modes radiate broadside with the two-lobed pattern of a short printed monopole working against its counterpoise, matching the asymmetric-dipole picture behind (4). Higher modes show nulls rotated about the ring axis, and their pattern orthogonality is what keeps the envelope correlation in Section 5 low. Feed position on the ring perimeter was chosen to sit on a current maximum of the broadside modes.

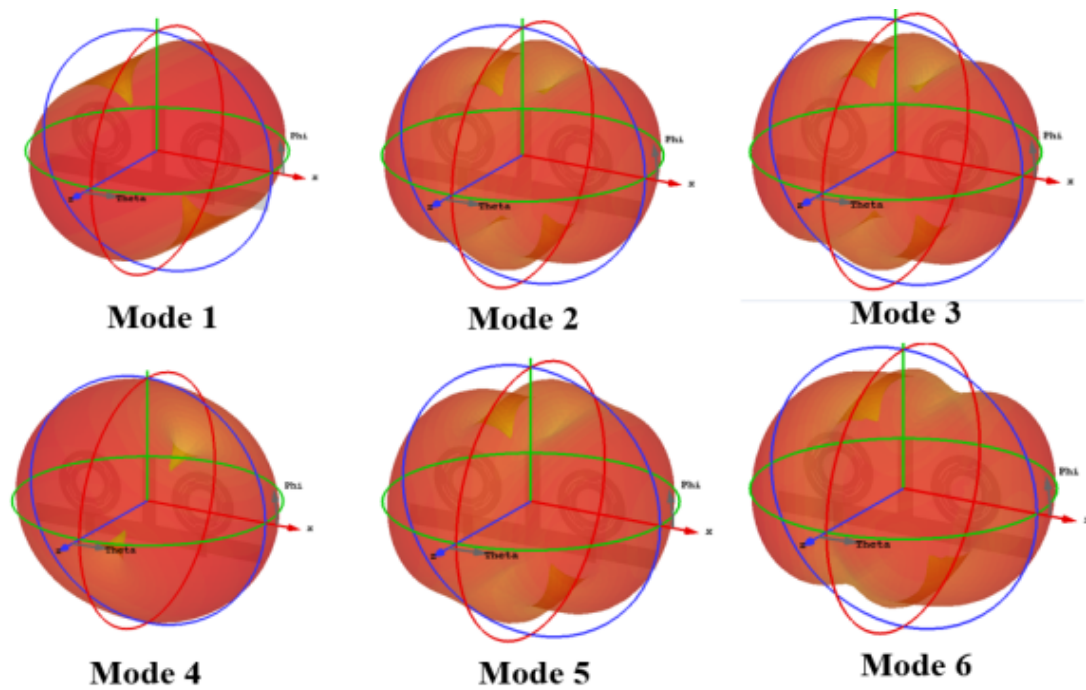


FIGURE 4. 3D far-field patterns of the proposed antenna at 11 GHz during various modes of CMA.

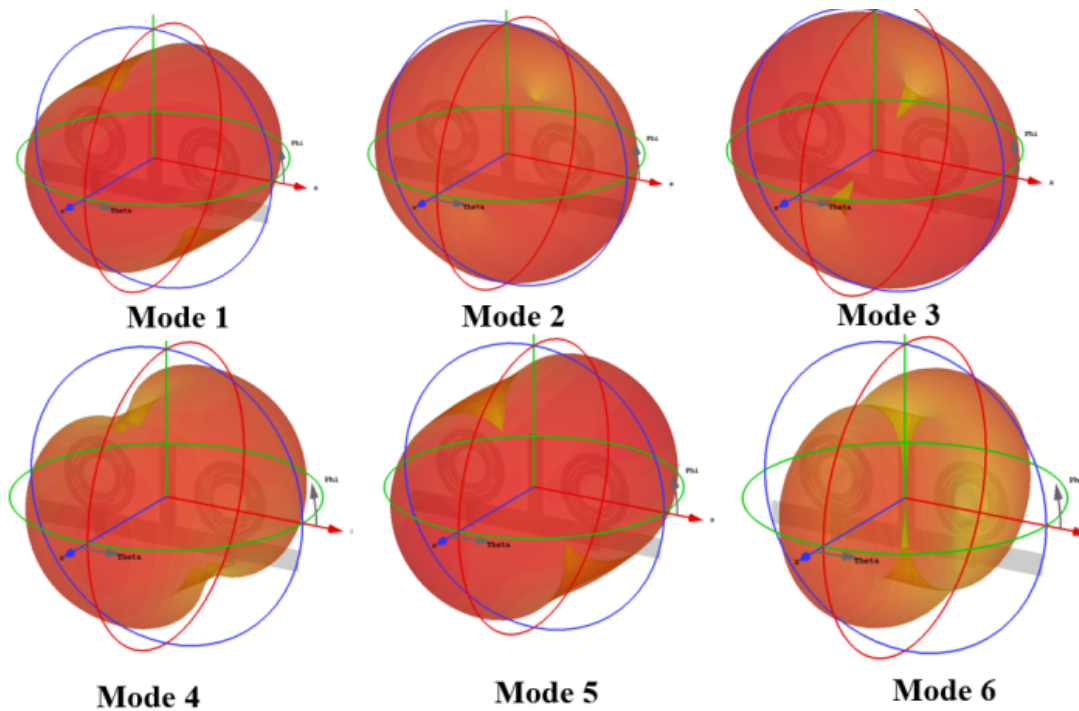


FIGURE 5. 3D far-field patterns of the proposed antenna at 8.85 GHz during various modes of CMA.

4. SIMULATED PERFORMANCE, DESIGN EVOLUTION AND PARAMETRIC STUDY

Full-wave simulation was carried out in Computer Simulation Technology (CST) Microwave Studio using the finite integration technique (FIT) time-domain solver, with an open (add-space) boundary applied on all faces and a $50\ \Omega$ discrete port 22 at each feed; the hexahedral mesh was adaptively refined until

the S -parameters converged between successive passes. Fig. 6 gives the four simulated S -parameters. The -10 dB band runs from 8.593 to 13.671 GHz, a width of 5.078 GHz and a fractional bandwidth of 45.6% about a centre of 11.13 GHz. The reflection coefficient reaches -20.8 dB at 11.36 GHz, and S_{11} and S_{22} lie on top of each other, as the mirror symmetry in Fig. 1 requires. Transmission stays below -20.7 dB everywhere in band and reaches -34.8 dB at 11.46 GHz.

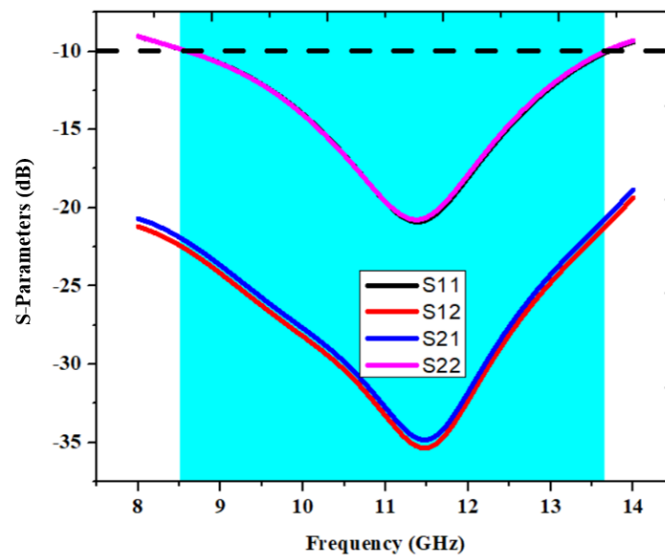


FIGURE 6. Simulated S -parameters of the proposed antenna.

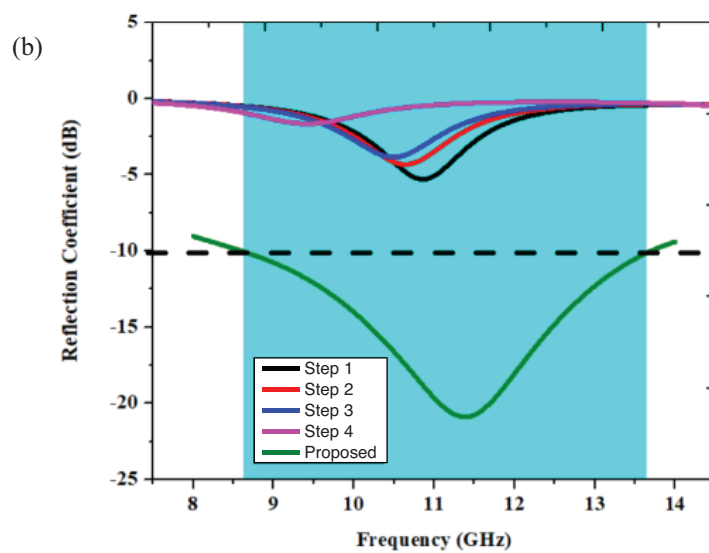
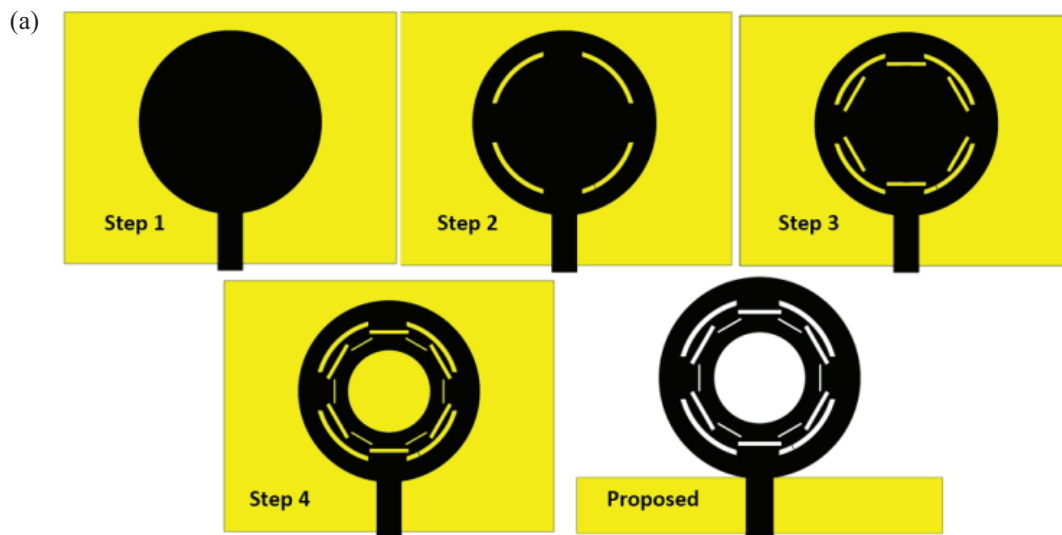


FIGURE 7. Design evolution of the proposed antenna. (a) Evolution steps and (b) reflection coefficient of each step.

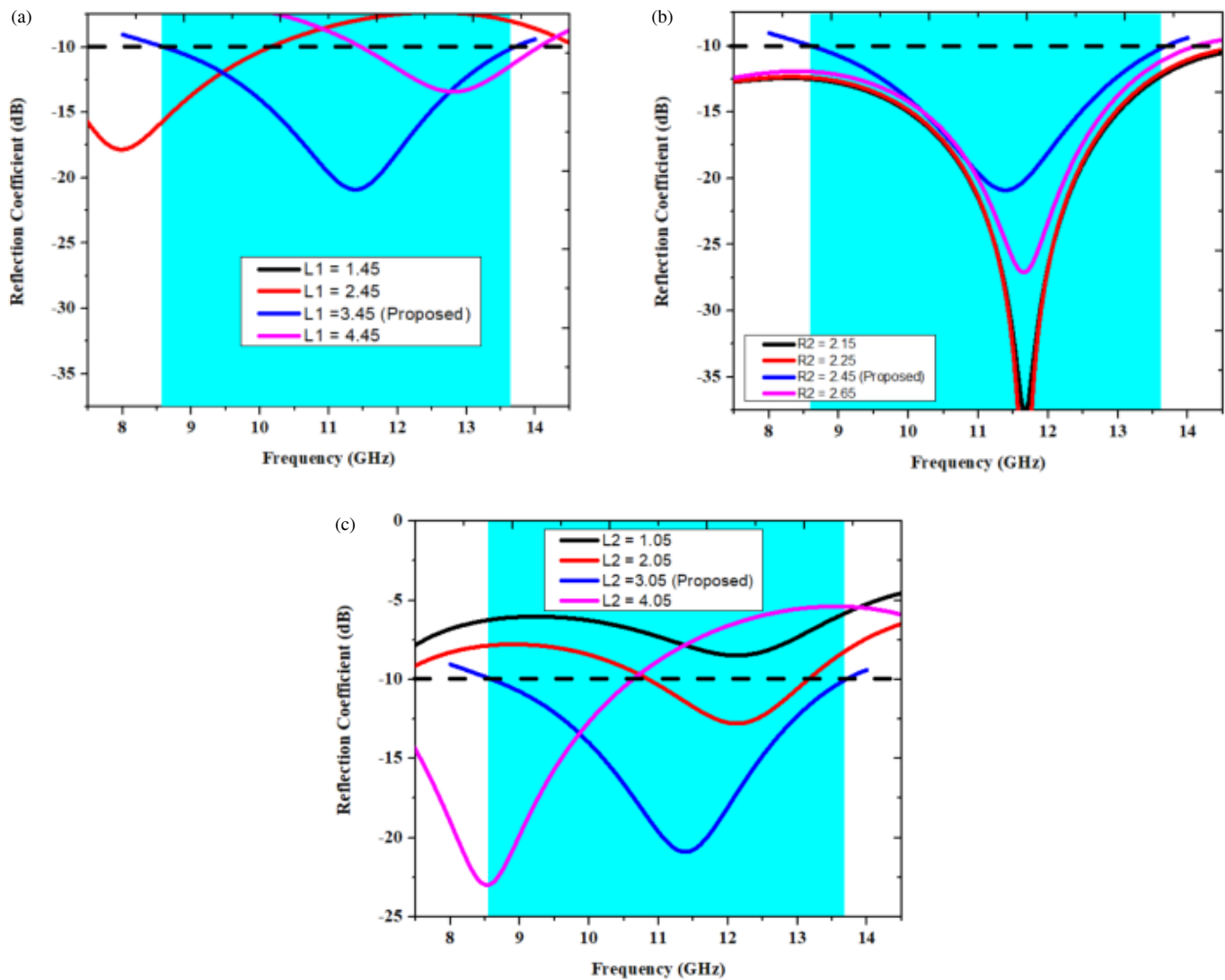


FIGURE 8. Parametric analysis of the proposed antenna when altering. (a) Feed length (L_1), (b) radius of the circular slot (R_2), and (c) ground plane length (L_2).

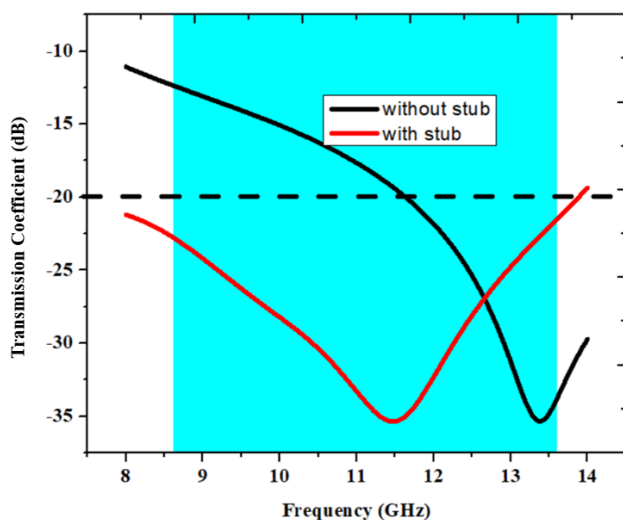


FIGURE 9. Transmission coefficient of the proposed antenna with and without centre stub.

Figure 7 traces the evolution. Step 1 is a plain circular disc on the fed line; Steps 2 and 3 add arc slots to the annulus; Step 4 opens the central aperture to form the ring; the final geometry carries the complete slot set of Fig. 1(b). None of the first four steps reaches a -10 dB match at any frequency in the swept range — Step 1 reaches only -4.9 dB near 11 GHz, and Steps 2 to 4 stay above -5 dB — whereas the final geometry produces the full 8.69–13.66 GHz response. The match is therefore a property of the completed structure, the slot set acting together with the truncated ground of Fig. 1(a), and not something the intermediate shapes achieve and the last step refines.

Figure 8 sweeps the three parameters that matter. Feed length L_1 controls the coupling into the ring. At 1.45 mm the line under-couples and the reflection coefficient sits near -7.7 dB across the band; at 2.45 mm a partial match appears with a minimum of -15.6 dB, but the response stays above -10 dB over part of the band; 3.45 mm gives the full 8.60–13.71 GHz band. Slot radius R_2 moves the depth and position of the mid-band minimum while leaving band edges nearly fixed, which is ex-

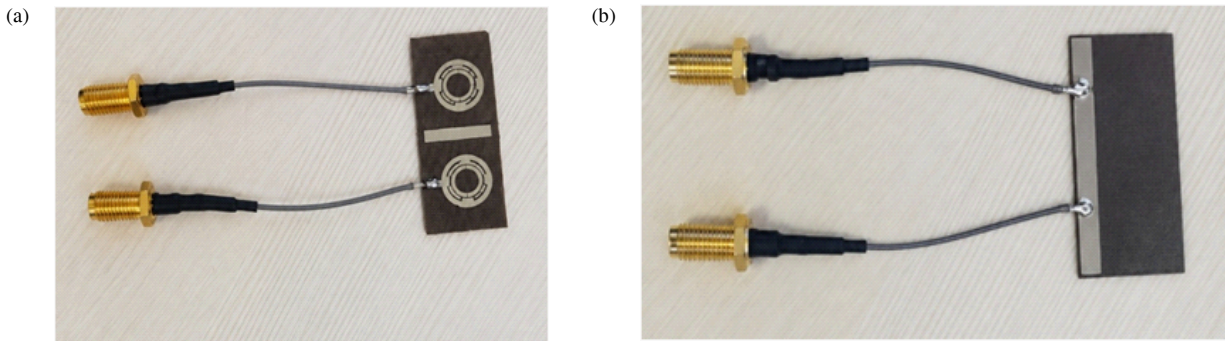


FIGURE 10. Fabricated prototype of the proposed antenna (a) front view and (b) back view.

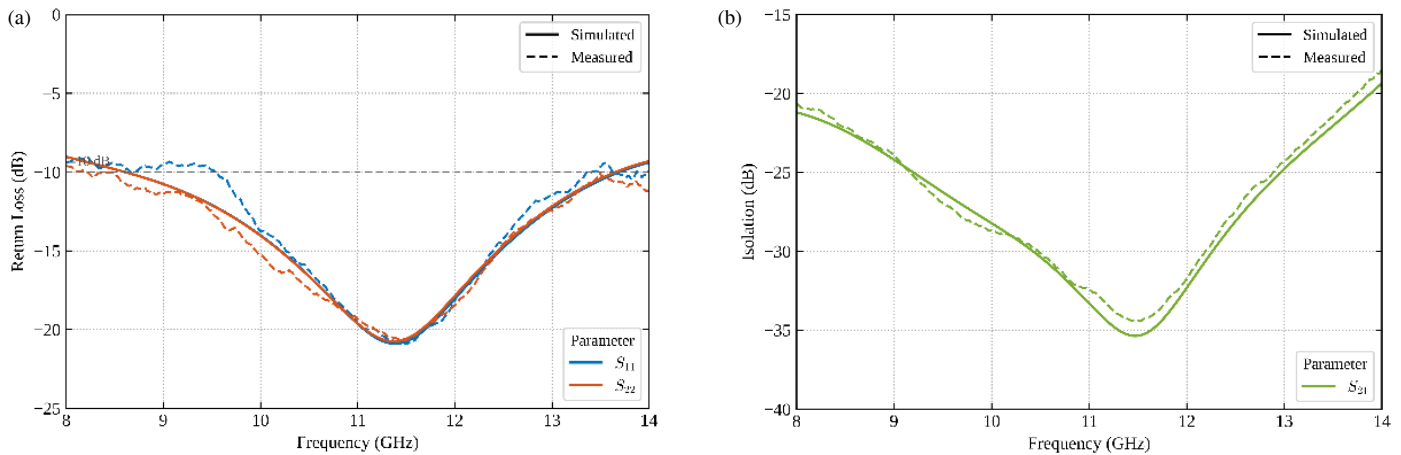


FIGURE 11. Measured S -parameters of the proposed antenna. (a) Reflection coefficient and (b) transmission coefficient.

pected because (7) makes the ring resonance depend on the sum $R_1 + R_2$, and R_2 is the smaller term; 2.45 mm was retained for the deepest in-band match consistent with the arc-slot geometry.

Ground length L_2 behaves exactly as (9) predicts. At 1.05 and 2.05 mm, the counterpoise is too small to support the asymmetric dipole mode of (4), and the response never matches — the in-band minimum is only -8.5 dB and -10.5 dB, respectively. At 3.05 mm the predicted quarter-wave edge is 14.43 GHz, and the simulated band closes at 13.73 GHz. At 4.05 mm the prediction drops to 11.45 GHz, and the swept curve indeed loses the upper band, degrading to -8.9 dB by 11 GHz and -5.6 dB by 13 GHz while remaining well matched below 10 GHz. A closed-form expression that tracks a four-point sweep in both directions is stronger evidence for the mechanism than the sweep alone.

Figure 9 isolates the contribution of the parasitic strip. Without it, in-band transmission degrades to -12.3 dB at worst. With it, worst-case in-band transmission improves to -20.7 dB and the peak deepens to -35.3 dB at 11.42 GHz. The strip therefore buys 8.4 dB of worst-case isolation and about 18 dB at the trap frequency — the trade is 2.4 mm of board width given to a non-radiating element, and no penalty in the impedance band, since Fig. 6 and Fig. 8 show the same edges with the strip present.

5. FABRICATION AND MEASURED RESULTS

A prototype was realised on 0.8 mm RT/Duroid 5880 and fed by edge-mounted SMA connectors. Fig. 10 shows the prototype layout. S -parameters were measured on a two-port vector network analyser with the unused port terminated in $50\ \Omega$; patterns, gain, and efficiency were measured in an anechoic chamber.

Figures 11(a) and (b) give the measured S -parameters. The measured -10 dB band runs from 8.00 to 13.96 GHz, a width of 5.96 GHz and a fractional bandwidth of 54.3%. The measured reflection minimum is -22.3 dB at 11.01 GHz. Measured transmission stays below -21.1 dB in band with a -33.4 dB peak near 11.06 GHz.

The measured band is 8.7% points wider than the simulated one, with almost all of it gained at the lower edge, which moves from 8.593 GHz to 8.00 GHz. Two effects account for the shift direction. The connector ground introduces additional counterpoise length beyond the modelled 3.05 mm, and (4) is sensitive to that length; substituting the measured edge into (4) implies an effective $L_1 + 2R_1$ of 14.81 mm rather than 14.65 mm, a difference of 0.16 mm that is within connector tolerance. Etch undercut on the arc slots would push in the same direction by lengthening the current path slightly beyond the nominal 9.1%. The upper edge lies within 2.1%, consistent with (9) depending on the ground geometry, which the connector does not alter.

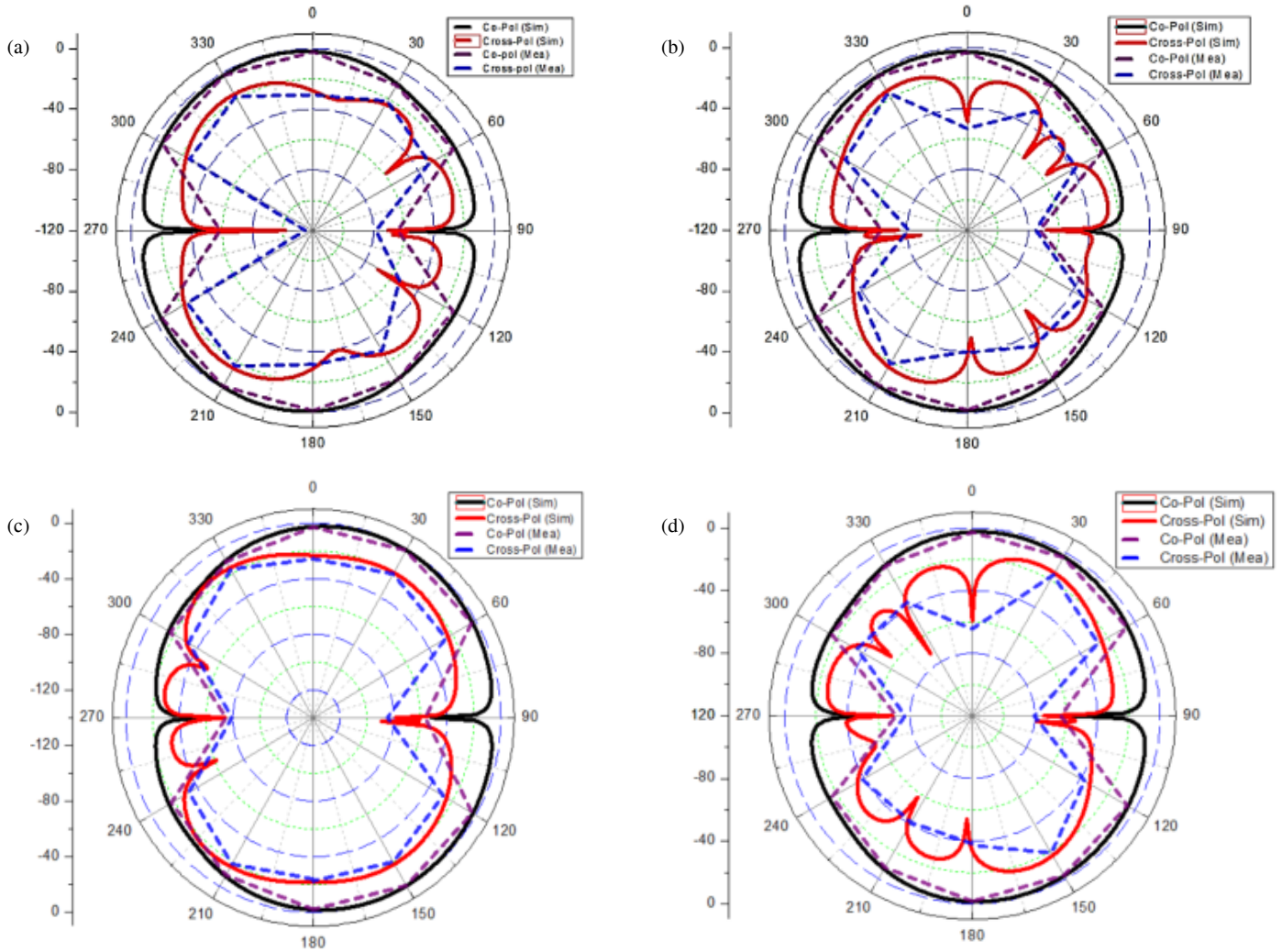


FIGURE 12. Simulated and measured far-field pattern of antenna 1 at (a) 10 GHz, (b) 12 GHz, antenna 2 (c) 10 GHz, and (d) 12 GHz.

Figure 12 compares simulated and measured co-polarised patterns for both elements at 10 and 12 GHz. Both elements radiate broadside with the two-lobed elevation pattern of a printed monopole over a truncated ground, and the two elements produce patterns that point away from each other in the plane containing them, with the geometric origin of the low correlation reported below. Agreement between simulation and measurement is close over the main lobe, with the usual divergence at the deepest nulls where chamber reflections and connector scattering dominate.

Figure 13 gives gain and radiation efficiency for both elements. Gain spans 4.0 to 6.5 dBi across the band, rising with frequency as the ring becomes electrically larger. Efficiency stays between 77% and 84%, consistent with a low-loss laminate and the resistive loading that the arc slots introduce. The two elements track each other closely, as symmetry requires.

Diversity performance was evaluated from measured S -parameters. For a two-port system, the envelope correlation coefficient, diversity gain, mean effective gain, channel capac-

ity loss, and ergodic capacity follow from

$$\rho_e = \frac{|S_{11}^* S_{12} + S_{21}^* S_{22}|^2}{(1 - |S_{11}|^2 - |S_{21}|^2)(1 - |S_{22}|^2 - |S_{12}|^2)} \quad (12)$$

$$\text{DG} = 10\sqrt{1 - |\rho_e|^2} \quad (13)$$

$$\text{MEG}_i = 0.5 \left(1 - \sum_{j=1}^2 |S_{ij}|^2 \right) \quad (14)$$

$$C_{\text{loss}} = -\log_2 \det(\Psi), \quad \Psi = \begin{bmatrix} \rho_{11} & \rho_{12} \\ \rho_{21} & \rho_{22} \end{bmatrix} \quad (15)$$

$$C = \log_2 \det \left(\mathbf{I} + \frac{\text{SNR}}{N} \mathbf{H} \mathbf{H}^H \right) \quad (16)$$

where $\rho_{ii} = 1 - \sum_j |S_{ij}|^2$ and $\rho_{ij} = -(S_{ii}^* S_{ij} + S_{ji}^* S_{jj})$.

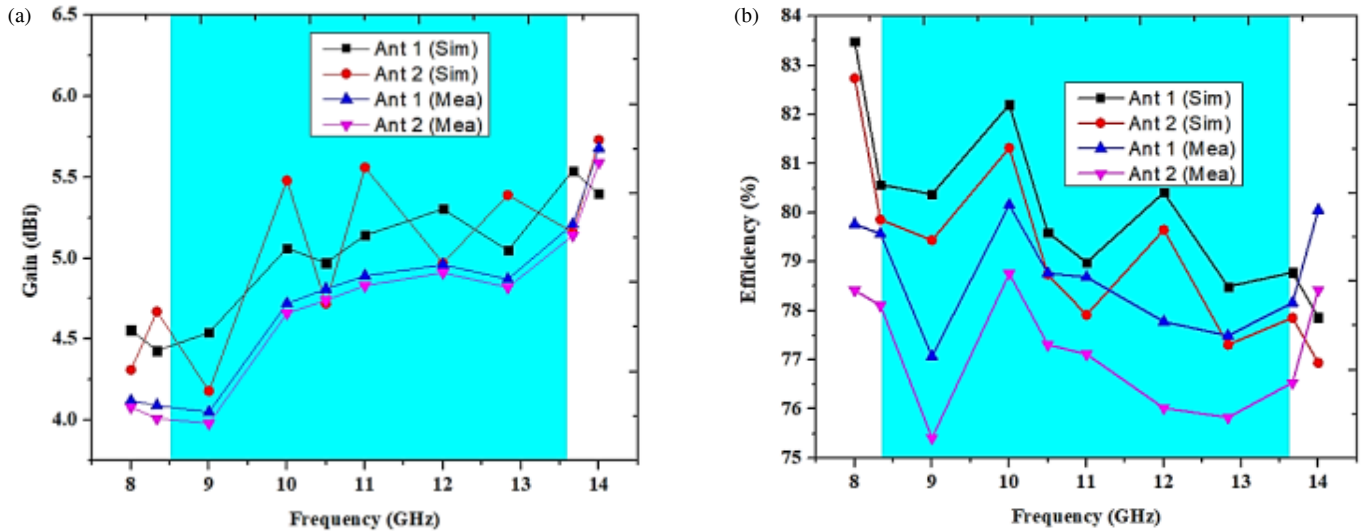


FIGURE 13. Simulated and measured. (a) Gain and (b) efficiency.

Figure 14 gives the results. Envelope correlation stays below 0.0038 in band, three orders of magnitude under the 0.5 limit normally quoted for diversity operation. Diversity gain remains above 9.981 dB against an ideal 10 dB. Mean effective gain lies between -3.05 and -3.52 dB for both elements, so the difference between them stays under 0.5 dB, and the branches are balanced. Channel capacity loss stays under 0.331 b/s/Hz, inside the 0.4 b/s/Hz limit, and the ergodic capacity of 13.01 to 13.28 b/s/Hz approaches the 2×2 ideal at the assumed signal-to-noise ratio. Simulated and measured curves overlay closely on every metric.

6. PHYSICALLY PARTITIONED TWO-PORT CIRCUIT MODEL

The model in Fig. 15 was built by walking the signal path from Port 1 inward and assigning one element to each geometric feature. No branch exists for fitting convenience.

The 68.68Ω line over the intact ground becomes a transmission-line section of length $L_2 = 3.05$ mm with the permittivity of (2); both values are fixed by (2) and (3) and not adjusted. The ground truncation appears as a shunt capacitance C_B at the ground edge. The exposed neck of length $L_1 - L_2 = 0.40$ mm carries no ground beneath it and becomes a series inductance fixed by the strip formula

$$L_N = 0.2\ell \left[\ln \frac{2\ell}{W_1} + 0.5 + 0.2235 \frac{W_1}{\ell} \right] = 56.8 \text{ pH} \quad (17)$$

with ℓ and W_1 in millimetres and L_N in nanohenries. Two series RLC branches shunt the radiating node: B_1 is the annular ring TM_{11} mode elongated by the arc slots, B_2 the next ring mode. They are in parallel because both current paths branch from the same point where the feed meets the ring. The central circular aperture resonates near 15.4 GHz with (7) applied to its perimeter, well above the band, and its fitted in-band reactance came out at zero — the aperture therefore contributes

to the model only through the ring geometry it defines, and no element is drawn for it.

Coupling between the elements travels through the shared board and appears as a series branch between two radiating nodes, split about a mid-node M . The parasitic strip shunts M through its own series RLC , so at the strip resonance the shunt admittance rises, the mid-node is loaded, and transmission between the ports falls. That is the circuit statement of the trap mechanism of Subsection 3.5.

Branches were parametrised by resonance and characteristic impedance rather than by L and C directly, which confines the search to geometries that can exist:

$$L_i = \frac{\pi Z_{0s,i}}{4\omega_i}, \quad C_i = \frac{1}{\omega_i^2 L_i} = \frac{4}{\pi Z_{0s,i}\omega_i} \quad (18)$$

The radiating admittance, node reduction, and port response are

$$Y_{rad} = \sum_{i \in \{1,2\}} \left(R_i + j\omega L_i + \frac{1}{j\omega C_i} \right)^{-1} \quad (19)$$

$$Y_{AA} = Y_{rad} + Y_m - \frac{Y_m^2}{Y_M}, \quad Y_{AB} = -\frac{Y_m^2}{Y_M} \quad (20)$$

with $Y_m = 2/Z_c$ and $Y_M = Y_{st} + 2Y_m$. Cascading the fixed feed network on both sides and converting the resulting chain matrix gives

$$S_{11} = \frac{A + B/Z_0 - CZ_0 - D}{A + B/Z_0 + CZ_0 + D}, \quad (21)$$

$$S_{21} = \frac{2}{A + B/Z_0 + CZ_0 + D}$$

Only the four branch triples and C_B were fitted by differential evolution, followed by a bounded least-squares polish, against the digitised full-wave S_{11} and S_{21} .

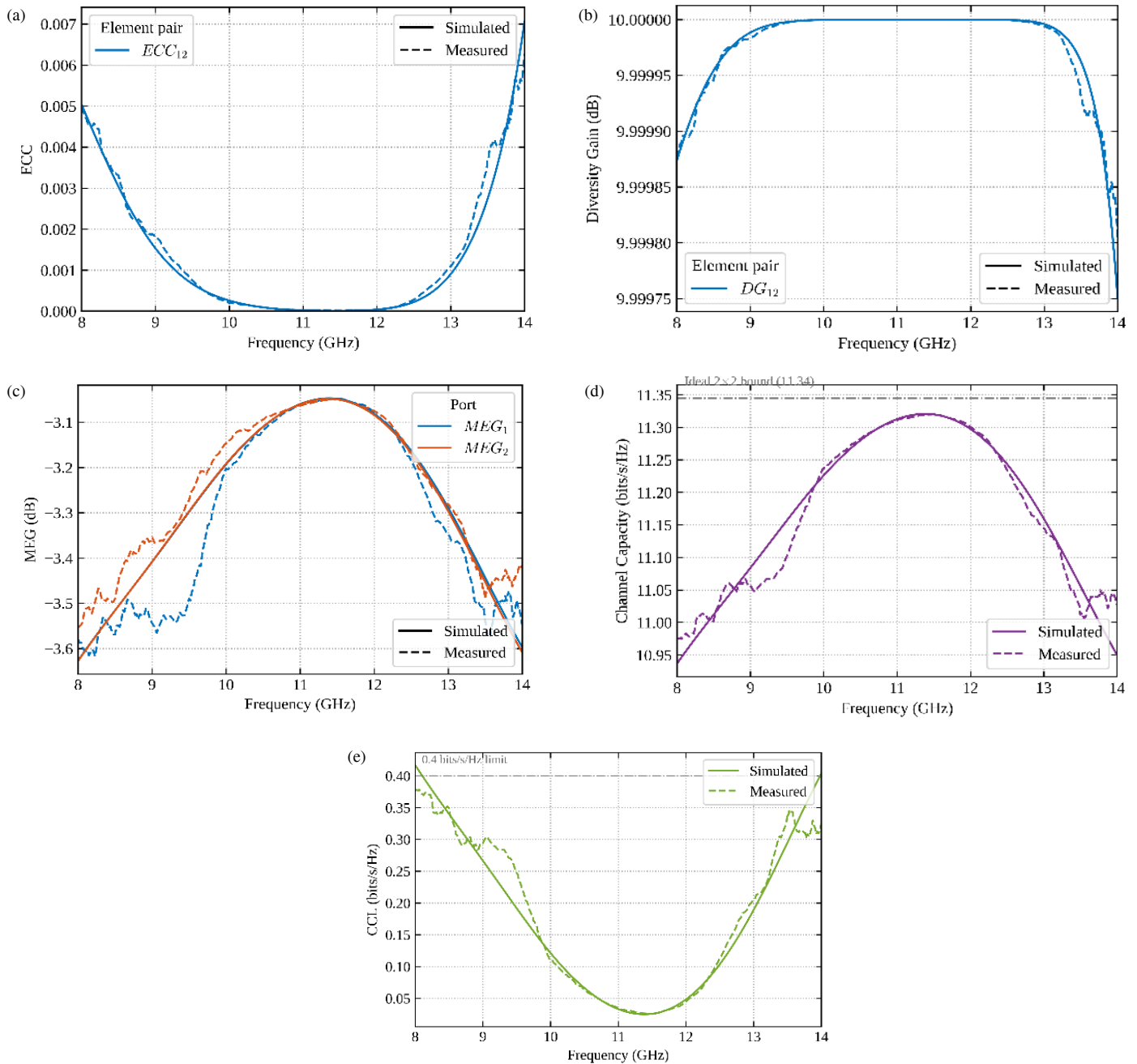


FIGURE 14. Simulated and measured MIMO metrics of the proposed antenna. (a) ECC, (b) diversity gain, (c) mean effective gain, (d) channel capacity, and (e) channel capacity loss.

Table 2 lists the extracted values with a validation column. The arc slots reduce the conducting track to about 0.883 mm, and the thin-strip estimate

$$Z_{0s} \approx \frac{\eta_0}{2\pi} \cdot \frac{\ln(2\ell/a) - 1}{\sqrt{\epsilon_{eff,0}}}, \quad a = \frac{w}{4} \quad (22)$$

gives $177.4\,\Omega$ for the B_1 arm and $137.2\,\Omega$ for the strip. The fitted values run 79% and 80% above those estimates. The offset is systematic rather than random and has a physical reading: (22) describes a strip radiating in isolation, whereas both features here terminate on a shared node and couple to the tracks

beside them, which raises the effective series reactance. A consistent offset across two independent features is explainable; a scattered one would not be.

Branch B_2 sits against its bounds, at 9.40 GHz and $460\,\Omega$. That is the honest limit of the partition: the simulated response has one broad minimum rather than two, so the second ring mode is only weakly separated, and the fit uses B_2 largely as a damping term. A five-branch version of this model reached 0.206 dB root mean square (RMS) on S_{11} , but every radiating impedance pinned at its ceiling with values 284% and 434% above (22), which explains nothing. The four-branch model

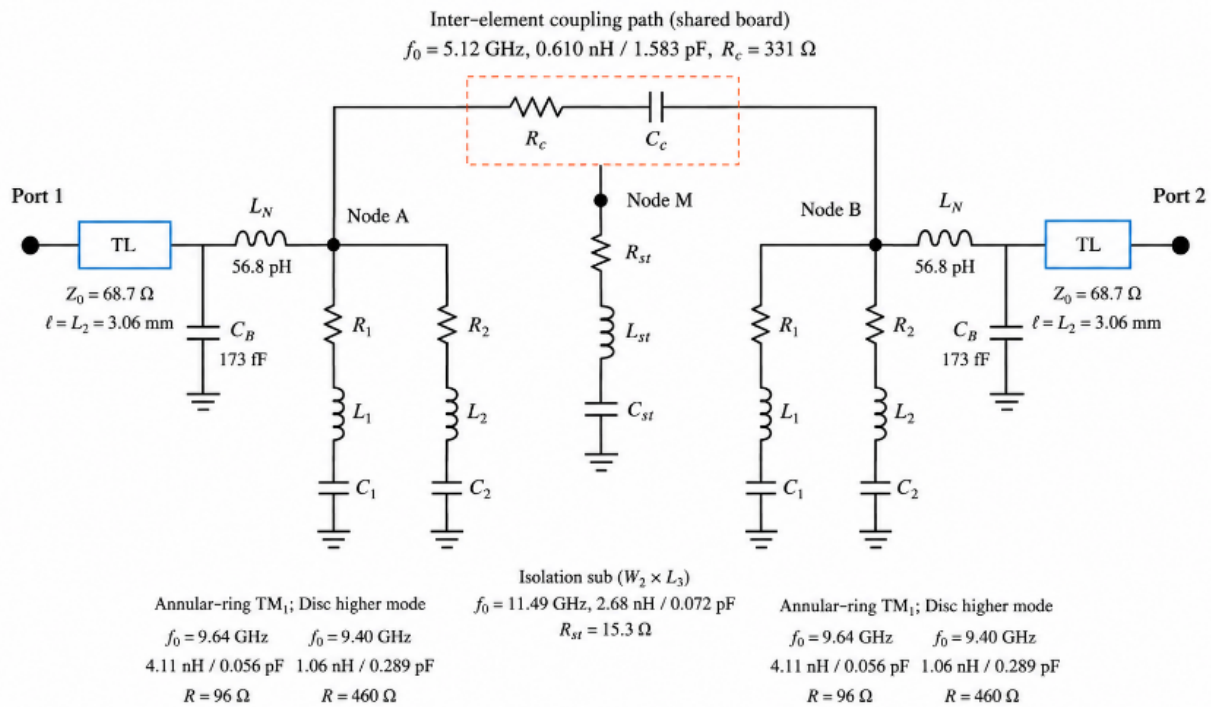


FIGURE 15. Physically partitioned two-port equivalent circuit of the proposed antenna. Fixed elements are computed from (2), (3), and (17); only the four series RLC branches and C_B are fitted.

TABLE 2. Extracted circuit elements, their physical origin, and validation.

Element	Physical feature	f_0 (GHz)	Z_{0s} (Ω)	R (Ω)	L (nH)	C (pF)	Validation
TL	68.7Ω line over intact ground, $\ell = L_2 = 3.06 \text{ mm}$	—	68.68	—	—	—	Fixed by (2), (3)
C_B	Ground-plane truncation fringing	—	—	—	—	0.173	Fitted, small
L_N	Exposed feed neck, $\ell = L_1 - L_2 = 0.40 \text{ mm}$	—	—	—	0.057	—	Fixed by (17)
B_1	Annular-ring TM_{11} , arc-slot elongated	9.642	316.9	95.9	4.109	0.066	(7) gives 9.372 GHz, 2.9%; (22) gives 177 Ω , +79%
B_2	Next ring mode	9.400	80.0	460.0	1.064	0.270	At bounds; weakly separated
Z_c	Coupling path through shared board	5.119	25.0	330.9	0.610	1.584	Below band, capacitive in band
Z_{st}	Parasitic strip, $W_2 \times L_3$	11.489	246.4	15.3	2.680	0.072	(10) gives 11.927 GHz, 3.8%; measured peak 11.43 GHz, 0.5%

was therefore preferred, at the cost of a 0.15 dB higher RMS error on S_{11} , because every element retains a physical meaning

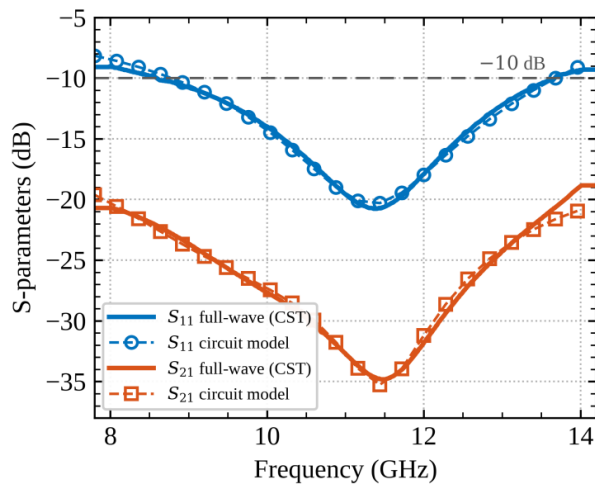
Figure 16 overlays the circuit response on the full-wave data. The model reproduces S_{11} with 0.352 dB RMS error and S_{21} with 0.611 dB. Band edges lie within 2.22% at the lower edge and 0.07% at the upper. The reflection minimum falls at -20.4 dB and 11.35 GHz against -20.8 dB and 11.36 GHz from the solver, and the isolation peak at -35.3 dB and 11.45 GHz against -34.8 dB and 11.46 GHz . Reproducing the isolation notch within 0.01 GHz from a model whose trap element was independently predicted by (10) is the result that justifies the partition.

7. COMPARISON WITH RECENT WORK

Table 3 compares the proposed antenna with recent multi-port designs. Two honest points must be made about it. The fractional bandwidth of 54.3% is lower than the ultra-wideband entries [5] and [2], which reach 124% and 74%; this follows from a band starting at 8.00 GHz rather than 3.0 or 8.5 GHz, with a far higher upper edge, not from a narrower match. Isolation of 21.1 dB is also below the 44 dB of [6] and the 25 dB of [5], both of which use orthogonal element placement or a defected ground — options this design deliberately avoids in order to keep the counterpoise intact and the outline narrow.

TABLE 3. Comparison with recently reported multi-port MIMO antennas.

Ref	Year	Ports	Substrate (ϵ_r , h mm)	Size (mm ²)	−10 dB band (GHz)	FBW	Isolation (dB)	CMA	Two-port ECM
[2]	2024	4	2.2 (0.5)	25 × 22	8.5–18.5	74.1%	> 15	No	No
[3]	2025	2	4.4 (1.6)	11 × 15	10.2–18.4	57.3%	> 15	No	Yes
[4]	2024	2	4.4 (1.6)	32 × 32	13.07–15.87	19.4%	n.s.	No	No
[5]	2025	2	2.2 (0.79)	54 × 30	3.0–13.0	124.4%	> 25	No	No
[6]	2025	4	2.2 (0.254)	35 × 35	24.0–40.0	50.0%	> 44	No	No
[20]	2023	1	4.4 (1.6)	47 × 46	3.22–12	115.37%	Single element	No	One-port only
This work	2026	2	2.2 (0.8)	40 × 15	8.00–13.96	54.3%	> 21.1	Yes	Yes

**FIGURE 16.** Full-wave and circuit-model S -parameters of the proposed antenna.

The advantages of the design lie in its low board height and its analysis. At 15 mm, the board is $0.40\lambda_o$ tall at the lower edge, giving a narrow strip outline that suits edge and conformal mounting, and it does so on $\epsilon_r = 2.2$, a low permittivity that offers no dielectric miniaturization. It is also the only entry in the table that carries both a modal analysis and a two-port circuit model validated against measured isolation; ref. [17] supplies the former and [20–22] the latter, never together and never on the same structure. Against the compact stub- and parasitic-strip-decoupled designs [2] and [3], the proposed 40×15 mm board is not the smallest in area, but it presents the lowest profile at $0.40\lambda_o$ on a low-permittivity substrate, while adding the joint modal and two-port circuit description those works omit.

8. CONCLUSION

This paper presents, analyses, and measures a two-port MIMO antenna for X- and Ku-band service. The radiator is a slotted annular ring on a $40 \times 15 \times 0.8$ mm³ RT/Duroid 5880 board, decoupled by a single parasitic strip on the radiating layer. Measurement gives an 8.00–13.96 GHz band, 54.3% fractional bandwidth, isolation better than 21.1 dB with a 33.4 dB peak, 4.0–6.5 dBi gain, and 77–84% efficiency, with envelope correlation under 0.0038 and diversity gain above 9.981 dB.

The design was reached analytically rather than by sweeping. A half-wave condition on the ring plus counterpoise predicts the lower edge to 1.11%; two independent routes to the ring resonance differ by only 4.14%, which quantifies the arc-slot path elongation at 9.1%; and a quarter-wave condition on the truncated ground predicts the upper edge and tracks a four-point sweep of the ground length in both directions. The parasitic strip worked as a three-quarter-wave trap, with its predicted resonance falling 4.3% from the measured isolation peak, and it provides 8.4 dB of worst-case isolation with no modification to the ground. A physically partitioned two-port circuit model, in which the feed line and neck inductance are fixed by closed-form expressions and every remaining element maps to a named feature, reproduces the full-wave response with 0.352 dB RMS error on S_{11} and 0.611 dB on S_{21} and locates the isolation notch to within 0.01 GHz.

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