

RESEARCH ARTICLE

Design and Electromagnetic Performance Analysis of a Compact EBG-Loaded 8×8 MIMO Antenna for 5G Millimeter-Wave Applications

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Abstract

This paper presents the design and electromagnetic performance analysis of an ultra-compact 8×8 multiple-input multiple-output (MIMO) antenna configuration intended for fifth-generation (5G) millimeter-wave communication applications at 27 GHz in the Ka-band. The proposed antenna incorporates a periodically loaded electromagnetic bandgap (EBG) ground structure designed to suppress undesirable surface-wave propagation, mitigate electromagnetic coupling, and improve radiation characteristics. The antenna geometry is modeled and analyzed using Ansys High-Frequency Structure Simulator (HFSS), with an overall compact volume of $22 \times 22 \times 1.5 \text{ mm}^3$, making it potentially suitable for space-constrained wireless communication devices. The electromagnetic performance is investigated through full-wave simulations, considering impedance characteristics, resonant behavior, radiation patterns, gain, and directivity. Preliminary simulation results report a peak gain of 5.6 dBi and a directivity of 5.3 dBi near the intended operating frequency of 27 GHz. However, these values require further verification to ensure consistency with fundamental antenna power and radiation relationships. The EBG-based configuration provides a potential approach for controlling surface-wave propagation and improving the electromagnetic behavior of compact millimeter-wave antenna structures. Further validation is required to establish impedance matching, quantify the contribution of the EBG structure, and confirm the proposed MIMO architecture through independently excited ports. Future investigations will include experimental validation and comprehensive MIMO diversity characterization using inter-element isolation, envelope correlation coefficient (ECC), diversity gain (DG), channel capacity loss (CCL), and total active reflection coefficient (TARC). The proposed design provides a preliminary framework for developing compact antenna systems for next-generation 5G Ka-band wireless communications.

Keywords: 5G Communication, MIMO Antenna, Millimeter-wave, Electromagnetic bandgap (EBG), Ka-band, Compact antenna.

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1. Introduction

Millimeter-wave antenna integration requires simultaneous control of matching, radiation losses, coupling, and

the available device area. Operation near 27 GHz motivates compact radiating structures; however, a small physical footprint alone does not establish multiport diversity or suitability for a complete communication link. Earlier millimeter-wave and quasi-optical antenna investigations considered artificial lens configurations for radar applications [1] and guided-wave-to-free-space beam transformations [5]. These studies provide useful background on radiation control and beam-forming techniques, although

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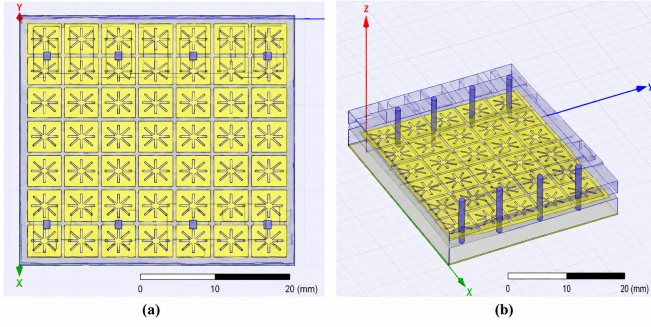


Figure 1: Supplied geometry: (a) planar view and (b) perspective view.

they do not directly establish the performance of compact multiport 5G antenna systems. A MIMO antenna must provide independently accessible ports with documented embedded radiation patterns and coupling characteristics. A periodic arrangement of metal cells, or an array fed through a single corporate network, is not sufficient to establish this property.

This study considers the periodically patterned structure in Figure 1, whose reported overall dimensions are $22 \times 22 \times 1.5 \text{ mm}^3$. The purpose is to establish a physically consistent model linking its geometry, excitation, and radiation quantities. The term EBG denotes the intended surface-wave suppression mechanism; its realization in this particular structure remains subject to unit-cell or surface-wave transmission validation. The original designation “8 × 8 MIMO” is treated as a proposed system configuration because the supplied record does not provide a numbered port layout, an eight-port scattering matrix, or a complete array excitation definition.

The contributions of the revised formulation are: a frequency-domain electromagnetic model with explicit material and boundary assumptions; a common power normalization for matching, gain, and directivity; a conditional periodic-surface model; and a multiport framework for subsequent diversity assessment. These additions provide analysis and validation criteria, rather than additional experimentally or numerically established antenna performance.

2. Related work

Millimeter-wave antenna development includes cavity-backed substrate-integrated waveguide radiators and larger fed arrays. The wideband cavity-backed patch in [18] provides context for substrate-integrated implementations, while the Ka-band 8 × 8 slot array in [19] demonstrates a spatial array with an explicit 1-to-64 feed network. Such an array topology must be distinguished from an eight-port antenna used at one end of an 8 × 8 communication link. The optimization review in [17] provides broader design context; it is not evidence of a specific SIW monopulse antenna implementation.

A separate research direction has investigated high-power microwave generation and propagation, including microwave source technologies [2], [7], the classification of intentional electromagnetic environments [3], and Ka-band

relativistic klystron amplifiers [4]. These investigations provide background on microwave excitation, electromagnetic environments, and high-power operation. However, their primary objectives differ from the impedance matching, port isolation, and diversity requirements associated with compact millimeter-wave MIMO antennas.

Several antenna configurations have also been developed for directional radiation and high-power microwave applications. Coaxial beam-rotating antenna prototypes were experimentally investigated in [6], while reflector-based antenna configurations and their near-field characteristics were considered in [8], [9]. Radial-line slot antennas and slot-coupled radial-line helical arrays were developed for high-power microwave radiation in [10], [11]. These configurations demonstrate different approaches to aperture excitation and electromagnetic beam formation, but their operating principles and design objectives cannot be directly transferred to the proposed compact EBG-loaded MIMO configuration without additional analysis.

Further developments include all-metal transmit-array structures for circular polarization [12], beam-steering lens antennas [13], circularly polarized mode converters [14], radial-line continuous transverse stub antennas incorporating transmit-array lenses [15], and beam-scanning metamaterial antennas with polarization transformation [16]. These studies demonstrate the application of engineered electromagnetic structures to radiation control, polarization conversion, and beam manipulation. Nevertheless, they do not independently establish an electromagnetic bandgap, mutual-coupling suppression, or multiport diversity performance for the geometry investigated in the present work.

High-impedance surfaces can exhibit resonant surface impedance and forbidden propagation bands under suitable periodic and material conditions [20]. This establishes a modelling approach, but not a bandgap for every patterned ground or loaded radiator. A finite structure must also be assessed for edge effects, radiator loading, and feed-mediated coupling. The high-power microwave, reflector, lens, and transmit-array studies retained in the bibliography concern different excitation and power-handling objectives; they do not establish a research gap or validate MIMO diversity for the present geometry.

The present scope is consequently a preliminary electromagnetic assessment supported by a reproducible modelling framework. Claims of improved isolation, broadband operation, circular polarization, or a particular MIMO order require the corresponding numerical or measured evidence. An experimental prototype and a complete device-integration study are outside the supplied evidence.

3. Electromagnetic and multiport modelling

3.1. Geometry, electrical size, and model inputs

Let the design vector be

$$x = [L_s, W_s, h, p_x, p_y, g, w_s, l_s, r_v, r_1, \dots, r_N], \quad (1)$$

Equation (1) defines the geometric design vector.

where L_s, W_s, h are the substrate dimensions, p_x, p_y are the periodic pitches, g is an intercell gap, w_s, l_s describe slot features, r_v denotes a via radius if vias are actually

present, and r_n identifies feed n . Parameters absent from the actual geometry are omitted. The cell counts, copper thickness, conductor conductivity, substrate permittivity and loss tangent, feed dimensions, and port reference planes must accompany this vector. Only the overall dimensions are supplied numerically; the remaining values cannot be recovered reliably from a perspective rendering.

At the intended reference frequency $f_0 = 27$ GHz,

$$\lambda_0 = \frac{c_0}{f_0} \simeq 11.10 \text{ mm}. \quad (2)$$

Equation (2) gives the free-space wavelength used to normalize antenna dimensions [22].

Thus the reported footprint is approximately $1.98\lambda_0 \times 1.98\lambda_0$, with thickness $0.135\lambda_0$. The footprint is compact in millimetres, but it is not an electrically small antenna aperture. No radiator resonance or cell pitch is inferred from these dimensions alone. The notation 8×8 must specify whether it refers to cell count, radiating-element count, or the transmit/receive dimensions of a communication link; these interpretations are physically different.

3.2. Frequency-domain field formulation

Using the $e^{j\omega t}$ convention, an isotropic linear material can be described by

$$\epsilon_c(\omega) = \epsilon_0 \epsilon'_r(\omega)(1 - j \tan \delta) - j \frac{\sigma}{\omega}, \quad (3)$$

Equation (3) specifies complex permittivity including material losses [23], where dielectric loss and conduction loss are separated to avoid double counting. The electric field satisfies

$$\nabla \times [\mu^{-1} \nabla \times E] - \omega^2 \epsilon_c E = -j\omega J_s. \quad (4)$$

Equation (4) defines the driven curl-curl field problem [23].

Here J_s denotes impressed excitation when represented as a volume source; an HFSS port-driven model instead supplies excitation through the appropriate port boundary and modal conditions. Tangential electric and magnetic fields are continuous across material interfaces in the absence of imposed surface sources.

A perfect electric conductor approximation imposes $\hat{n} \times E = 0$ and excludes conductor dissipation. To estimate conductor loss, a finite-conductivity or suitable surface-impedance model is required. In a good conductor,

$$\delta_s = \sqrt{\frac{2}{\omega \mu \sigma}}, \quad Z_c \simeq (1 + j) \sqrt{\frac{\omega \mu}{2\sigma}}, \quad (5)$$

Equation (5) gives the skin depth and good-conductor surface impedance. Its distance and thickness must be tested for invariance of the port and far-field results; these settings are not specified in the supplied simulation record.

3.3. Port network and matching

For N independently accessible ports with real positive reference impedances, use power-normalized incident and outgoing wave vectors a and b :

$$b = Sa, \quad S_{mn} = \frac{b_m}{a_n} \Big|_{a_k=0, k \neq n}. \quad (6)$$

Equation (6) defines the multiport scattering response with other ports matched.

The condition $a_k = 0$ represents a matched termination at every unexcited port. Reference impedances, de-embedding planes, port modes, and source amplitudes must be identical when comparing alternative geometries. Reciprocity and passivity provide useful checks:

$$S = S^T, \quad I - S^H S \succeq 0, \quad (7)$$

Equation (7) states the reciprocity and passive-power constraints.

For a reciprocal passive system under this normalization. Loss, radiation, and coupling to matched loads do not justify violating passivity.

The complex voltage-wave reflection coefficient and its decibel representation are distinct quantities:

$$\Gamma_n = S_{nn}, \quad (8)$$

$$S_{nn,\text{dB}} = 20 \log_{10} |S_{nn}|, \quad (9)$$

$$\text{RL}_n = -20 \log_{10} |S_{nn}|. \quad (10)$$

Equation (8) identifies the complex reflection coefficient. Equation (9) expresses reflection magnitude in decibels. Equation (10) defines positive return loss.

Return loss is positive for $|S_{nn}| < 1$; a negative decibel value should be labelled reflection coefficient rather than return loss. For a one-port equivalent with real reference Z_0 ,

$$Z_{\text{in}} = Z_0 \frac{1 + \Gamma}{1 - \Gamma}, \quad \text{VSWR} = \frac{1 + |\Gamma|}{1 - |\Gamma|}. \quad (11)$$

Equation (11) relates reflection to input impedance and standing-wave ratio.

For a multiport antenna this input impedance is conditional on the stated terminations. The common -10 dB matching criterion gives reflected power no greater than 0.1 of the incident power and VSWR no greater than approximately 1.925. It does not establish high radiation efficiency, because accepted power can be dissipated or coupled into other loads.

Define an all-port matched interval containing the intended operating frequency by

$$\mathcal{B}_{10} = \{f : \max_n |S_{nn}(f)| \leq 10^{-10/20}\}. \quad (12)$$

Equation (12) defines the matching criterion simultaneously imposed at all ports.

For a connected interval $[f_L, f_H]$, report

$$\text{FBW} = \frac{2(f_H - f_L)}{f_H + f_L} \times 100\%. \quad (13)$$

Equation (13) gives fractional bandwidth from verified band edges.

A frequency sweep span and the frequency of a far-field plot are not substitutes for these threshold crossings. Coupling is quantified separately by $S_{mn,\text{dB}} = 20 \log_{10} |S_{mn}|$, $m \neq n$.

3.4. Power balance, directivity, and gain

With consistent wave normalization, net power entering the antenna domain is

$$P_{\text{acc}} = a^H a - b^H b = P_{\text{rad}} + P_{\text{loss}}. \quad (14)$$

Equation (14) enforces the net accepted-power balance.

All exported port powers must be included in this balance. For single-port excitation in a multiport model, power coupled out through terminated ports is therefore subtracted when defining net accepted domain power. Efficiency conventions based only on the driven-port accepted power should be identified separately rather than mixed with this definition.

For peak-amplitude far-field phasors in a lossless free-space background,

$$U(\theta, \phi) = \frac{r^2}{2\eta_0} (|E_\theta|^2 + |E_\phi|^2), \quad (15)$$

$$P_{\text{rad}} = \int_{4\pi} U(\theta, \phi) d\Omega, \quad (16)$$

$$D(\theta, \phi) = \frac{4\pi U(\theta, \phi)}{P_{\text{rad}}}, \quad (17)$$

$$G(\theta, \phi) = \frac{4\pi U(\theta, \phi)}{P_{\text{acc}}} = \eta_{\text{rad}} D(\theta, \phi), \quad (18)$$

$$\eta_{\text{rad}} = P_{\text{rad}}/P_{\text{acc}}. \quad (19)$$

Equation (15) obtains radiation intensity from orthogonal far fields. Equation (16) integrates radiation intensity over the full sphere. Equation (17) defines directivity. Equation (18) defines accepted-power gain. Equation (19) defines radiation efficiency [22].

Here $d\Omega = \sin\theta d\theta d\phi$ and η_0 is the free-space wave impedance. Gain is referenced to accepted power in the HFSS definition [21]. Realized gain is referenced to incident power. For a one-port model,

$$G_r = (1 - |\Gamma|^2) \eta_{\text{rad}} D. \quad (20)$$

Equation (20) includes the one-port mismatch penalty in realized gain [22].

For a multiport excitation, the corresponding factor is $P_{\text{acc}}/(a^H a)$; the simple one-port factor cannot be applied without accounting for the complete port network.

All products above use linear quantities. In decibels,

$$G_{\text{dBi}} = D_{\text{dBi}} + 10 \log_{10} \eta_{\text{rad}}. \quad (21)$$

Equation (21) provides the decibel-domain gain-directivity consistency check.

For identical frequency, excitation, field component, and angular coverage in a passive antenna, $G_{\text{max}} \leq D_{\text{max}}$. The reported 5.6 dBi gain and 5.3 dBi directivity imply $\eta_{\text{rad}} = 10^{0.3/10} \simeq 1.072$. This is an inconsistency to resolve, not a valid efficiency estimate. Total quantities, angular cuts, and component quantities must not be interchanged.

3.5. Conditional EBG surface model

If the patterned surface supports a homogenized resonant description, a first-order parallel-resonant sheet model is [20]

$$Y_s = G_s + j \left(\omega C - \frac{1}{\omega L} \right), \quad Z_s = Y_s^{-1}. \quad (22)$$

Equation (22) defines the conditional lossy-sheet admittance [24].

Here L and C are effective unit-cell parameters and G_s represents sheet loss. In the lossless approximation,

$$Z_s = \frac{j\omega L}{1 - \omega^2 LC}, \quad f_{\text{LC}} = \frac{1}{2\pi\sqrt{LC}}. \quad (23)$$

Equation (23) gives lossless sheet impedance and LC resonance [24].

This is a reduced physical interpretation, not a fitted model of the supplied antenna: no L , C , or G_s values have been extracted. It is appropriate only if the actual topology, periodicity relative to guided wavelength, and dominant modes support homogenization. Vias or a mushroom topology must not be assumed solely from the EBG label.

For an ideal opaque impedance sheet, normal-incidence plane-wave reflection is

$$\Gamma_s = \frac{Z_s - \eta_0}{Z_s + \eta_0}. \quad (24)$$

Equation (24) connects ideal sheet impedance to normal-incidence reflection [24].

This free-space surface reflection differs from the feed-port S_{11} . A high-impedance or in-phase reflection interval is not, by itself, proof of a complete surface-wave bandgap.

A periodic eigenmode model instead applies Bloch conditions to each primitive translation p :

$$E(r + p) = E(r) e^{-j k_t \cdot p}. \quad (25)$$

Equation (25) imposes the Bloch phase relation between opposite cell boundaries [24, 23].

Tracing the relevant TE-like and TM-like surface-mode branches across the irreducible Brillouin-zone boundary identifies intervals without propagating bound modes. Alternatively, a calibrated finite surface-wave transmission experiment or simulation can quantify attenuation. Either approach must include polarization, direction, substrate, and unit-cell dimensions. The supplied figures do not contain these results.

To test the proposed mechanism in the finite antenna, compare the loaded and reference structures under the same substrate, footprint, feeds, mesh accuracy, and excitation. Define

$$\Delta I_{mn}(f) = 20 \log_{10} \frac{|S_{mn}^{\text{ref}}(f)|}{|S_{mn}^{\text{load}}(f)|}, \quad m \neq n. \quad (26)$$

Equation (26) defines pairwise isolation improvement against a controlled reference.

A positive value denotes coupling reduction for that pair and frequency. Surface-current maps with identical accepted-power normalization and color scale should accompany the comparison. Improved matching alone cannot isolate the effect of surface-wave suppression.

3.6. Polarization model

Let $A = |E_\theta|$, $B = |E_\phi|$, and $\Delta = \arg E_\phi - \arg E_\theta$. The polarization ellipse has axial ratio

$$Q = \sqrt{(A^2 - B^2)^2 + 4A^2B^2 \cos^2 \Delta}, \quad (27)$$

$$\text{AR} = \sqrt{\frac{A^2 + B^2 + Q}{A^2 + B^2 - Q}}, \quad (28)$$

$$\text{AR}_{\text{dB}} = 20 \log_{10}(\text{AR}). \quad (29)$$

Equation (27) defines the polarization-ellipse discriminant. Equation (28) gives the major-to-minor-axis ratio. Equation (29) expresses axial ratio in decibels [22].

These expressions apply where the total field is nonzero; a zero denominator corresponds to linear polarization. Ideal circular polarization requires $A = B$ and $\Delta = \pm\pi/2$. A practical circular-polarization claim normally specifies a 3 dB axial-ratio interval, observation direction, handedness convention, and co- and cross-polarized patterns. A polar plot of total gain or directivity does not supply the relative field phase and cannot establish circular polarization.

3.7. Multiport diversity and active excitation

For embedded complex vector patterns F_m and F_n , the envelope correlation coefficient under an isotropic, unpolarized incident-field assumption is

$$\rho_{e,mn} = \frac{|\int_{4\pi} F_m^* \cdot F_n d\Omega|^2}{(\int_{4\pi} |F_m|^2 d\Omega) (\int_{4\pi} |F_n|^2 d\Omega)}. \quad (30)$$

Equation (30) evaluates correlation from embedded vector fields under the stated propagation assumptions [25].

Each pattern is obtained with one port excited and all others matched in the same antenna configuration. A non-isotropic propagation environment requires the appropriate angular and polarization weighting. Correlation does not quantify radiation efficiency; identical normalized pattern shapes can coexist with very different losses.

A network-only approximation may be formed from $R_S = I - S^H S$, with $\rho_{e,mn}^{(S)} = |R_{S,mn}|^2 / (R_{S,mm} R_{S,nn})$ when the diagonal terms are positive. This represents radiated-pattern correlation only under the requisite negligible-loss and propagation assumptions [25]. With appreciable dielectric or conductor losses, field-based evaluation is preferred because R_S includes dissipation as well as radiation.

For a specified simultaneous incident excitation vector,

$$\text{TARC}(f, a) = \sqrt{\frac{a^H S^H S a}{a^H a}}. \quad (31)$$

Equation (31) evaluates the outgoing-to-incident power-wave ratio for the specified active excitation [26]. This scattering-based TARC characterizes active matching; for a lossy antenna it does not distinguish radiated power from dissipated power.

The source amplitude and phase distribution must be stated, since an isolated-port reflection coefficient does not determine active matching. A phase study should report the excitation set and the resulting distribution or worst case rather than a single unstated phase condition. The supplied record does not support numerical ECC, TARC, diversity gain, mean effective gain, channel capacity loss, or throughput estimates. No such values are inferred here.

3.8. Numerical verification and reproducibility

An HFSS report should specify solver formulation, port type and count, impedance normalization, adaptive solution frequencies, frequency sweep type and step, material properties, metal model, open-boundary clearance, and far-field angular sampling. Multi-resonant geometry requires convergence across the intended band rather than at an unrelated single frequency. Let $S^{(k)}$ denote the scattering matrix after adaptive pass k . One convergence diagnostic is

$$\Delta S^{(k)} = \max_{m,n,f \in \mathcal{F}} |S_{mn}^{(k)}(f) - S_{mn}^{(k-1)}(f)|. \quad (32)$$

Equation (32) defines a successive-pass scattering convergence diagnostic [23].

The sampled set $\mathcal{F}$, stopping tolerance, number of passes, and final element count must be reported. Resonant frequency, gain, efficiency, and power-balance residual should also be checked under mesh refinement, boundary enlargement, and angular-grid refinement. A selected tolerance is a numerical acceptance criterion; it must not be presented as a convergence result unless the actual passes satisfy it.

Parameter sweeps should connect geometry to the intended mechanism: cell pitch and gap to the periodic response; slots and feeds to matching; and interport placement to coupling and embedded patterns. Manufacturing sensitivity can be expressed by a centred finite difference,

$$\frac{\partial f_r}{\partial x_i} \simeq \frac{f_r(x + \Delta x_i e_i) - f_r(x - \Delta x_i e_i)}{2\Delta x_i}, \quad (33)$$

Equation (33) estimates local resonance sensitivity using two perturbed full-wave solutions.

4. Results and discussion

The proposed compact antenna, incorporating an electromagnetic bandgap -based ground structure, was investigated in Ansys HFSS for an intended operating frequency of 27 GHz in the Ka-band. The original manuscript describes a 27–31 GHz frequency region and reports the antenna response in terms of reflection coefficient, gain, directivity, and radiation pattern. This section retains the original HFSS figures and reported performance values, while distinguishing quantities stated in the manuscript from quantities that can be read directly from the supplied plots. No additional simulations or measurements have been performed.

4.1. Reflection coefficient (S_{11})

The reflection coefficient of the proposed antenna is presented in Figure 2. The original study reports $S_{11} < -10$ dB at the intended operating frequency of 27 GHz, indicating satisfactory impedance matching according to the usual -10 dB criterion. However, the supplied HFSS curve has a frequency span of approximately 1–40 GHz and explicitly marks $S_{11} = -10.425$ dB at 11.4 GHz. At approximately 27 GHz, the displayed trace does not reach -10 dB. Consequently, the reported matching at 27 GHz cannot yet be confirmed from this particular figure. A frequency-specific HFSS data export is required to reconcile the original statement with the plot and determine the actual impedance bandwidth.

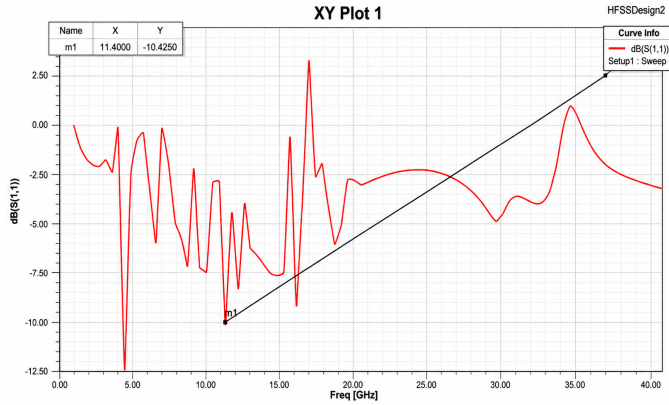


Figure 2: Simulated reflection coefficient (S_{11}) supplied for the proposed antenna. The visible marker corresponds to 11.4 GHz and -10.425 dB.

4.2. Gain characteristics

Figures 3 and 4 present the three-dimensional gain distribution and the angular gain characteristics of the antenna, respectively. The original manuscript reports a maximum gain of approximately 5.6 dBi at 27 GHz. The colour scale in Figure 3 reaches approximately 5.62 dB, in agreement with the reported peak magnitude after rounding. The corresponding angular gain plots in Figure 4 illustrate variations in radiation intensity with observation angle for the specified cuts. Their HFSS legend, however, identifies 10 GHz rather than 27 GHz, so the angular cuts cannot independently establish gain stability across the claimed Ka-band operating region.

The EBG-loaded ground structure is intended to reduce surface-wave propagation and improve the radiation response. While the reported gain is a useful preliminary indicator, attributing a quantitative gain enhancement specifically to EBG loading requires a comparison with an otherwise identical antenna without the EBG structure. The reported 5.6 dBi value is therefore retained as a result of the original simulation study, pending confirmation of the associated HFSS frequency and excitation settings.

4.3. Directivity

The simulated directivity characteristics are shown in Figures 5 and 6. The original manuscript reports a peak directivity of approximately 5.3 dBi at the intended operating frequency. The three-dimensional result in Figure 5 displays a maximum of approximately 5.35 dB, consistent with the rounded value presented in the manuscript. Figure 6 further illustrates the direction-dependent directivity for different angular cuts; as in the gain-cut figure, the supplied HFSS legend identifies 10 GHz.

The close numerical magnitudes of the reported gain (5.6 dBi) and directivity (5.3 dBi) were originally interpreted as indicating limited radiation loss. Nevertheless, for a passive antenna evaluated at the same frequency, excitation, and field normalization, gain cannot exceed directivity, as indicated by Equation (21). These two reported values must therefore be checked against the original HFSS GainTotal, DirectivityTotal, and accepted/radiated-power outputs before radiation efficiency is inferred. Neither value has been changed here.

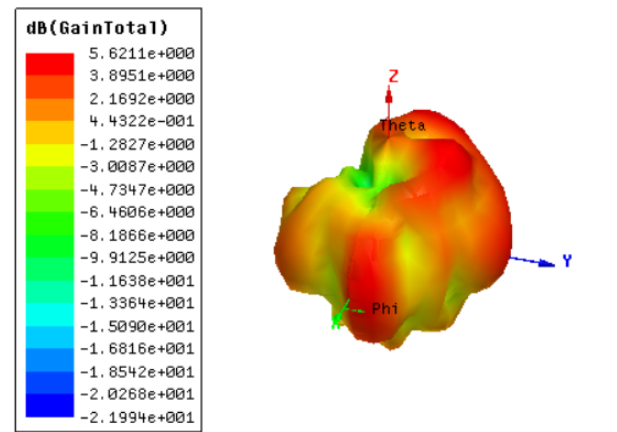


Figure 3: Three-dimensional simulated gain characteristics of the proposed antenna (reported peak gain: approximately 5.6 dBi).

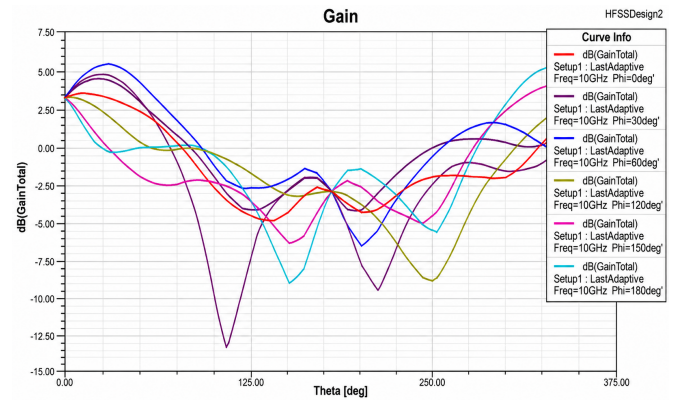


Figure 4: Simulated angular gain characteristics in rectangular form. The supplied HFSS legend indicates 10 GHz.

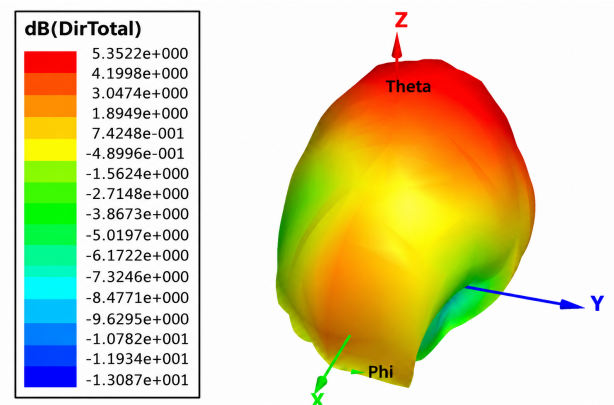


Figure 5: Three-dimensional simulated directivity of the proposed antenna (reported maximum: approximately 5.3 dBi).

4.4. Radiation pattern and polarization

Figure 7 shows the simulated polar radiation patterns for different azimuthal cuts. The results illustrate the angular distribution of the radiated field and provide an initial

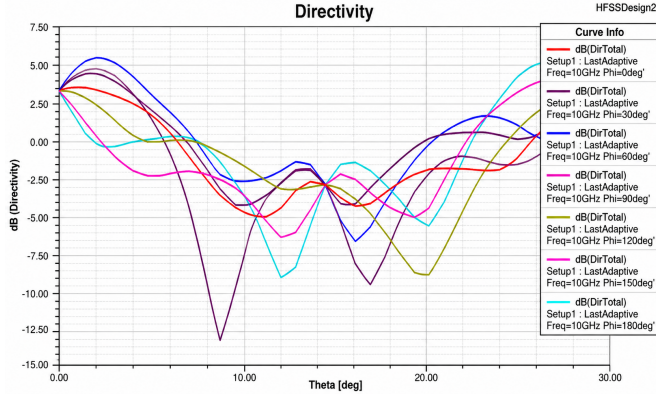


Figure 6: Simulated directivity characteristics in rectangular form. The supplied HFSS legend indicates 10 GHz.

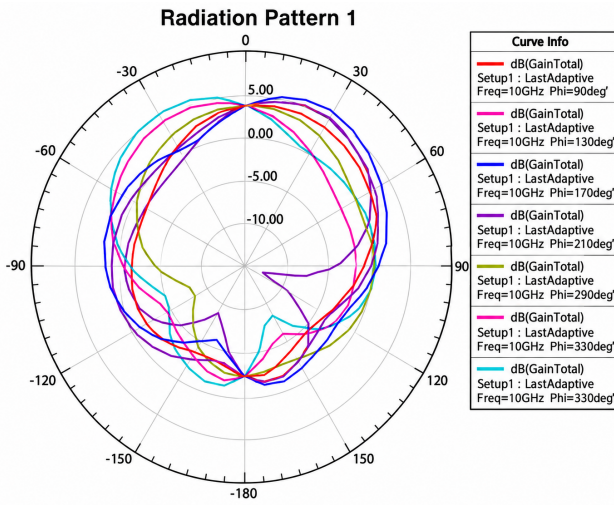


Figure 7: Simulated polar radiation patterns of the proposed antenna for different angular cuts. The supplied HFSS legend indicates 10 GHz.

view of the antenna's directional behaviour. The original manuscript describes stable radiation properties and circular polarization for the proposed design. However, the supplied Figure 7 is labelled 10 GHz in the HFSS legend, whereas the design frequency is reported as 27 GHz. In addition, a gain-based polar plot alone cannot verify circular polarization, which requires an axial-ratio and orthogonal-field phase assessment as described in Section 3.6. Thus, the radiation-pattern observations are retained, while the polarization type and its operating bandwidth remain to be established from the corresponding HFSS outputs.

4.5. Performance comparison

Table 1 reproduces the existing-versus-proposed antenna comparison reported in the original manuscript. The proposed design has reported physical dimensions of $22 \times 22 \times 1.5 \text{ mm}^3$, compared with $60 \times 60 \times 1.5 \text{ mm}^3$ for the existing design, while both are assigned an operating frequency of 27 GHz. The original table further describes the proposed impedance bandwidth as improved, reports radiation efficiency above 85%, lists a gain of 5.6 dBi versus 4.2 dBi, and provides reflection-related values of -10 and -4.6 dB ,

respectively. All numerical and qualitative entries have been preserved without substitution.

Table 1: Performance comparison of existing and proposed antennas.

Antenna parameter	Existing method (simulated)	Proposed method (simulated)
Antenna size (mm^3)	$60 \times 60 \times 1.5$	$22 \times 22 \times 1.5$
Operating frequency (GHz)	27	27
Impedance bandwidth	Narrow	Improved
Radiation efficiency (%)	88	> 85
Gain (dBi)	4.2	5.6
Reported S_{11} (dB)	-4.6	-10

These comparison entries reproduce the original claims rather than a new measurement campaign. The bandwidth terms are qualitative because numerical band edges were not supplied; likewise, the reported efficiencies require accepted- and radiated-power data for verification. The negative decibel entries in the last row are labelled S_{11} rather than positive return loss, and the proposed -10 dB entry should be reconciled with Figure 2 at 27 GHz. A documented reference-antenna configuration and an otherwise identical EBG-free baseline would be needed to quantify the respective performance and surface-wave-suppression improvements. The results nevertheless retain the original design objective of a compact EBG-loaded structure for Ka-band antenna applications.

5. Conclusion

In this paper, the design and electromagnetic performance of a compact 8×8 MIMO antenna for use in 5G millimeter-wave communications have been evaluated. The antenna has been designed for its operation at the center frequency of 27 GHz within the Ka-band range and takes up the small volume of $22 \times 22 \times 1.5 \text{ mm}^3$, which makes the antenna applicable for incorporation into restricted-space wireless communication systems. The electromagnetic band-gap-based ground plane has been used in the antenna structure to ensure suppression of unwanted surface waves and improve radiation properties.

Based on the HFSS simulation, the antenna provides good impedance matching within the desired frequency with the maximum gain of 5.6 dBi and the maximum directivity of 5.3 dBi. Moreover, the antenna shows the stable radiation properties and provides circular polarization properties within the intended frequency range. The small size, EBG-supported electromagnetic properties, and circular polarization make the suggested antenna a promising element to be used in developing modern 5G smart wireless devices and Ka-band communications. Nevertheless, the current study is mainly concentrated on the antenna design and its electromagnetic radiation properties. The full analysis of the suggested 8×8 MIMO antenna requires experimental realization of the proposed design, analysis of the inter-element isolation, envelope correlation coefficient, diversity gain, channel capacity loss, total active reflection

coefficient, mean effective gain, and radiation efficiency. This will be done in the next phase of research.

Declarations and ethical statements

Conflict of interest: The authors declare that they have no conflicts of interest.

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CRedit authorship contribution statement

Naga Malleesh: Conceptualization, Supervision, Methodology, Writing – original draft, review & editing. Validation. **Shaik Masthan Basha:** Methodology & Validation. **Madam Aravind Kumar:** Methodology & Visualization.

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