Isolation Improvement of MIMO Antenna Using Novel EBG and Hair-Pin Shaped DGS at 5G Millimeter Wave Band

SOUMIK DEY, (Student Member, IEEE), SUKOMAL DEY, (Senior Member, IEEE), AND SHIBAN K. KOUL, (Life Fellow, IEEE)

1Department of Electrical Engineering, Indian Institute of Technology Palakkad, Palakkad, Kerala 678557, India
2Centre for Applied Research in Electronics, Indian Institute of Technology Delhi, New Delhi 110016, India

Corresponding author: Sukomal Dey (sukomal.iitpkd@gmail.com)

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ABSTRACT This paper proposes a hybrid decoupling method based on a novel electromagnetic bandgap (EBG) structure and hair-pin shaped defected ground structure (DGS) to obtain high isolation between 2-element multiple input multiple output (MIMO) antenna at 5G millimeter wave band over 27.5 – 28.35 GHz. The proposed EBG designed on stacked dielectric substrates, achieves a wide frequency band-gap between 26.2 – 32.03 GHz (20%). A 2 × 3 array of the EBG is arranged between two electromagnetically coupled radiating patches in order to suppress the surface wave coupling. Substrate integrated waveguide (SIW) feeding network and cavity are strategically incorporated in the antenna design for improving the radiation performance and minimizing the losses from the feed. EBG shows an average isolation improvement of 13.9 dB within 5G band as compared to unloaded MIMO antenna. The additional reduction in coupling is achieved by placing hair-pin DGS (HP-DGS) on the ground plane, resulting in maximum isolation improvement of 47.7 dB at 27.94 GHz. The prototype of the MIMO was fabricated and experimentally verified. Measured peak isolation between the antennas is obtained as 71.9 dB, having a gain of 9 dBi and front to back ratio (FTBR) of 19.8 dB. A good diversity performance is also noticed for the designed MIMO with envelope correlation coefficient (ECC) of 0.00015, diversity gain (DG) of 9.99, and channel capacity loss (CCL) of 0.025 bits/Hz/sec. Later, SIW corporate feed network is designed for 4-element linear array loaded with EBG and HP-DGS to achieve higher gain and narrow beamwidth. The array was fabricated and the measured results are found in good accordance with the simulation results. The peak gain, beamwidth, and FTBR of the array are 13.3 dBi, 16.2°, and 19.97 dB respectively.

INDEX TERMS Antenna array, defected ground structure (DGS), electromagnetic bandgap (EBG), mutual coupling reduction, multiple input multiple output (MIMO).

I. INTRODUCTION
Inception of the fifth-generation (5G) wireless technology to fulfill the demands of efficient and high-speed transmission throughput requires multiple antennas at the transmitting and receiving ends [1], [2]. To meet the requirement of space constraint in modern internet of things (IoT) based wireless gadgets, antenna elements need to be densely packed. In common system circuit boards, the existence of surface wave at the air-dielectric interface gives rise to cross-talk between the closely spaced antennas. This deteriorates the channel capacity and efficiency of the MIMO system [3] and also causes scan blindness in phased array [4]. Different methods to reduce the crosstalk in MIMO/phased array can be listed as coupled line resonator [5], neutralization line [6], [7], polarization isolator [8], capacitively-loaded loop (CLL) [9], metamaterial superstrate [10]–[12], split ring resonator (SRR) [13], [14], EBG [15]–[22], DGS [23], [24], and metamaterial absorber [25]. The mutual coupling problem becomes more severe at the millimeter (mm) wave transceiver system. Thick electrical thickness of the substrate at mm-wave originates the high-order surface wave modes [26], [27]. In addition to that, space wave coupling between the array elements enhances correlation and hence
reduces the scanning range and gain of the beam steerable antenna [28]. Till to date, only few attempts were made to address the coupling reduction in mm-wave circuits [29]–[34]. With the growing interest in mm-wave communication, coupling needs to be controlled in the MIMO antenna with a minimum increase in the design complexity. In similar works reported earlier especially at mm-wave range, methods for improving the impedance matching and radiation efficiency have not been considered in the design process. In this paper, a compact uniplanar EBG is designed following a detailed analysis based on the equivalent circuit model (ECM). A hybrid isolator based on the EBG and HP-DGS is introduced for isolation improvement between two H-plane coupled MIMO antennas. Simulation of the MIMO antenna and the EBG is performed using High-Frequency Structure Simulator (HFSS) software. Later experimental results are found consistent with the simulation results. A 4-element array loaded with EBG and HP-DGS is designed that has features of high gain, low side lobe level (SLL), and high FTBR. The salient features of the present work are (i) unique design strategy of the EBG on stacked dielectric substrates with different relative permittivities, causing high attenuation of the coupling field over a wide band (ii) EBG-DGS based hybrid isolation method that achieves maximum isolation improvement of 47.7 dB within mm-wave range (iii) new variation of the mm-wave MIMO antenna and 4-element linear array which integrates the SIW cavity along with proximity feeding method to obtain simultaneously wideband matching, high FTBR and limit the propagation of surface wave. Sections II describes the geometry of the proposed uniplanar EBG and ECM. The single antenna element and its simulation results are presented in section III. Schematic of the decoupled MIMO antenna and its isolation characteristics are discussed in section IV. Section V presents fabricated MIMO antenna and its measurement results, followed by state-of-the-art performance comparison. Design of SIW cavity-backed 4-element linear array is discussed in section VI. Finally, the conclusion of the work is drawn in section VII.

II. DESIGN AND ANALYSIS OF PROPOSED EBG

A. UNIT CELL GEOMETRY

EBG is a 2D periodic structure which generally classified into two types—mushroom and uniplanar based on the unit cell geometry. The second kind of configuration has the advantage of without having shorting pins which simplifies the design. EBG supports different surface wave modes with each one of the modes having different field configurations. The mode with zero cutoff frequency is called transverse magnetic (TM) mode. EBG suppresses the propagation of the surface wave within its frequency bandgap due to its high impedance surface nature and also exhibits in-phase reflection similar to a perfect magnetic conductor (PMC) surface [35]. Fig. 1(a) shows the unit cell geometry of the proposed uniplanar EBG consists of two metal layers separated by two substrates with different dielectric constants \((\varepsilon_r)\). The bottom metal layer acts as a ground plane. Two substrates are Taconic TLY-5 \((\varepsilon_{r1} = 2.2)\) and FR-4 \((\varepsilon_{r2} = 4.4)\) with loss tangent values of 0.0009 and 0.02 respectively. The two substrates have thicknesses \((h_1)\) and \((h_2)\) of 0.51 mm and 0.8 mm. Compared with the earlier reported EBGs, the novelty of the proposed geometry is the use of two substrates with different dielectric constants. This causes multiple reflections of the surface wave at the interface between the top and bottom layers due to a change in refractive index \((n)\). Here \(n \propto \sqrt{\mu_r \varepsilon_r}\) with \(\mu_r\) denotes the relative permeability with a value of 1 for the non-magnetic material. Multiple reflections may also cause destructive interference inside stacked dielectric substrates, resulting into additional attenuation of surface wave over broad frequency range. This leads to wide bandgap of the EBG. Stacking of two substrates can be considered as a composite dielectric (CD) layer having a total thickness of 1.31 mm. The effective dielectric constant \((\varepsilon_{eff})\) of the CD layer is found as 3.45 using (1)

\[
\sqrt{\varepsilon_{eff}} = \left(\frac{\sqrt{\varepsilon_{r1}h_1} + \sqrt{\varepsilon_{r2}h_2}}{h_1 + h_2}\right)
\]

B. DISPERSION CHARACTERISTICS

The surface wave propagation characteristic of the EBG is obtained from the dispersion diagram which is a graphical representation of wavenumber \((k)\) against frequency \((f)\). For a lossy medium like EBG, \(k\) exhibits a nonlinear variation with frequency. The numerical simulation of the EBG is performed using the Eigen modes solver of the HFSS imposing periodic unit cell boundary, as shown in Fig. 1 (b).
height of the radiation box is assumed five times the total thickness of the unit cell. Periodic master-slave boundaries are assigned at the walls of the radiation box along x and y axes. While the wall along the + z-axis is terminated with a perfectly matched layer (PML) boundary. The 90° rotational symmetry in the unit cell geometry allows solving the Eigen frequencies only over the Brillouin zone triangle (Γ-X-M). Fig. 2 shows the dispersion characteristic of the proposed EBG that exhibits a frequency bandgap from 26.2 to 32.03 GHz (5.83 GHz or 20 %) between the two lowest order propagating modes. This bandgap is observed within the region covered by the light line which separates the unguided plane wave region and the guided surface wave region. The wideband bandgap of the proposed EBG is attributed to the inter-cell connection between adjacent unit cells. This enhances the total inductance \(L\) and increases the bandgap bandwidth (BW) which is proportional to \(\sqrt{L/C}\). Here \(C\) denotes the overall capacitance of the unit cell. Final dimensions of the proposed EBG are \(P = 2.05, w = 0.15, g = 0.15, S = 0.25, R_i = 0.525, R_o = 0.825, t = 0.035\) mm. The unit cell has periodicity of 0.19 \(\lambda_0\), where \(\lambda_0\) is the free space wavelength at the bandgap center frequency of 29.12 GHz.

**C. PARAMETRIC ANALYSIS**

The geometry of the unit cell consists of a center cross dipole whose arms are convoluted in the form of sectoral patches. Fours diagonal slits having length ‘S’ are included within the sectoral patches to control the BW of the frequency bandgap. The sensitivity of the frequency bandgap on the unit cell dimensions is further analyzed using parametric variations, shown in Fig. 3. Fig. 3 (a) depicts the dispersion plot against different radius \(R_i\) of the cross dipole over the region of \(0\) to \(X\) with phase variation (Phase X) from 0° to 180°. The increase in \(R_i\) causes the increase in periodicity of the unit cell and shifts the two modes towards lower frequency. The bandgap between the two modes is invariant with change in \(R_i\), because the \(L\) and \(C\) increase in same proportions, keeping the \(L/C\) ratio same. Variation of the diagonal slit length (S) provides the bandwidth control of the EBG as illustrated in Fig. 3(b). The increase in \(S\) enhances the slot capacitance of the top metasurface. This in effect reduces the Eigen frequency of the second mode. This frequency corresponds to zero group velocity \((d\omega/d\beta = 0)\) and determines the upper limit of the EBG bandgap. On contrary, the change in \(S\) does not affect the wave dispersion of the first mode. The bandgap of the proposed EBG decreases with an increase in \(S\) due to an increase in capacitance. The fractional bandwidth (FBW) of the bandgap changes from 23 % to 8.24 % when \(S\) varies from 0.15 to 0.45 mm.
TABLE 1. Comparison of the proposed EBG with previous reported works.

| Ref. | Center Freq. (GHz) | Bandgap (GHz FBW) | Periodicity (λ0) | Thickness (λ0) | Type |
|------|-------------------|------------------|-----------------|---------------|------|
| [16] | 5.63              | 0.58 (10.3 %)    | 0.1             | 0.029         | Mushroom |
| [18] | 5.125             | 0.45 (8.78 %)    | 0.143           | 0.017         | Uniplanar |
| [19] | 6.472             | 1.153 (17.8 %)   | 0.324           | 0.032         | Uniplanar |
| [29] | 61.95             | 6.3 (10.17 %)    | 0.047×0.202     | 0.103         | Uniplanar |
| Proposed EBG | 29.12             | 5.83 (20 %)      | 0.19            | 0.127         | Uniplanar |

\[ R_{p1} = 1.024Z_0 \tanh \left( 2.025 \frac{w_m}{h} \right) \]  \hspace{1cm} (6)

Here \( h = h_1 + h_2 \) is the total profile height. The gap capacitance \( C_{g1} \) is determined from even and odd mode capacitances \( (C_e \text{ and } C_0) \) using (7) [36]

\[ C_{g1} = 0.5C_0 - 0.25C_e \]  \hspace{1cm} (7)

The weak coupling between the widely spaced TLs causes a small value of \( C_e \). The estimated values of circuit parameters are: \( L_e = 1.98 \text{ nH}, L_{p1} = 2.06 \text{ nH}, C_{p1} = 0.0366 \text{ pF}, C_{p2} = 0.0359 \text{ pF}, C_{g1} = 0.0034 \text{ pF and } R_{p1} = 29.18 \Omega \). In the ECM, the EBG unit cells are configured as a T network with a series branch consisting of a parallel resonant circuit \( (L_{e1} - C_{e1}) \), and shunt branch is represented by a series resonance \( (L_{e2} - C_{e2}) \). In the series resonance inductance \( L_{e2} \) occurs due to induced current on the bottom ground plane, forming a current loop between top and bottom layers. This current couples with the magnetic field of the surface wave. The fringing field at the edges of slots in the top metasurface and the overlapping area between the two metal layers contribute to the series and shunt capacitances in the ECM. The circuit parameter values of EBG are obtained in the Advanced design system (ADS) using the curve fitting technique. The final values are: \( L_{e1} = 3.82 \text{ nH}, L_{e2} = 10.25 \text{ nH}, C_{e1} = 0.0087 \text{ pF}, C_{e2} = 0.073 \text{ pF} \). Fig. 5 (a) shows \( S_{21} \) between the two open-end TLs without and with EBG loading in the middle. The \( S_{21} \) of ECM is obtained in ADS and the result shows a close match with HFSS simulation. Over frequency bandgap of the EBG, \( S_{21} \) is significantly attenuated with maximum suppression centers at 27.5 GHz. To illustrate the advantage of the stacked dielectric substrates, the open ended TLs with and without EBG loading are simulated on CD material having height of 1.31 mm and uniform dielectric constant 3.45. The magnitude of \( S_{21} \) is suppressed by 24.2 dB more for EBG loaded open end TL on the stacked dielectric materials as compared to the same on CD layer.

Effective medium parameters of the EBG are also found from the scattering parameters. The medium permittivity (\( \varepsilon \)) and permeability (\( \mu \)) are related to refractive index \( n \) and impedance \( Z \) by equations \( \varepsilon = n Z \) and \( \mu = n Z \) where \( Z \) and \( n \) are obtained from \( S_{11} \) and \( S_{21} \) using (8) and (9) [37]

\[
Z = \sqrt{\frac{(1 + S_{11})^2 - S_{21}^2}{(1 - S_{11})^2 - S_{21}^2}}
\]  \hspace{1cm} (8)
Fig. 5 (b) shows the medium parameters of the EBG in which real component of $\varepsilon$ is negative from 23.72–30.36 GHz. The real part of the refractive index $n = \sqrt{\varepsilon \mu}$ are found to be close to zero over the 5G operating band. This confirms that the designed EBG behaves as zero index metamaterial.

Performance of the proposed EBG is compared with the reported mushroom and uniplanar EBGs, shown in Table 1. The unit cell in [29] also operates in mm-wave frequency, but the EBG in this paper possesses wider bandwidth. Besides asymmetric unit cell geometry of [29], causes the bandgap to be polarization sensitive.

FIGURE 6. Geometry of the proposed SIW cavity-loaded proximity feed patch antenna with T-shaped stub for impedance matching.

III. SIW CAVITY LOADED PATCH ANTENNA

A. ANTENNA GEOMETRY

Fig. 6 presents the schematic of the patch antenna loaded with SIW cavity. The radiating square patch of the antenna having length $L_p$ is excited with an electromagnetically coupled or proximity feeding method. Here instead of directly providing the signal excitation at the microstrip feed, a SIW based transition is made at the input section. The reason for this feeding arrangement is explained later. The microstrip feed with the SIW transition is printed on Taconic TLY-5 having a thickness of 0.51 mm. The substrate FR-4 with a height of 0.8 mm is used for radiating square patch and the SIW cavity with a rectangular aperture. Three sides of the cavity are enclosed by periodically arranged shorting pins which are perforated through the two substrates. Via diameter is used as 0.4 mm with a periodicity of 1 mm. A pair of ‘T’ shaped stubs is placed symmetrically on both sides of the microstrip feed to achieve broadband impedance matching. The length and width of the stubs are determined through a parametric analysis in HFSS. Radiating patch of the antenna is located at the open end of the microstrip feed. The wave radiates from the open end of the microstrip line and capacitively couples with the top radiating patch. This capacitance causes impedance mismatch of the antenna, which is balanced through the inductance provided by the T-shaped stubs. The two stubs are located at a distance of 0.3 mm from the open end of the microstrip line. The final design parameters are $L_x = 25$, $L_y = 15$, $L_p = 1.8$, $w_1 = 1.1$, $g_e = 0.35$, $w_c = 0.96$. All units are in millimeters.

FIGURE 7. (a) Equivalent circuit of the proximity feed antenna unloaded with SIW cavity (b) simulated reflection coefficient.

B. EQUIVALENT CIRCUIT AND SIMULATION RESULTS

The equivalent circuit of the proposed antenna is shown in Fig. 7 (a). The proximity feed patch without SIW cavity is modelled here to keep the analysis simple. In the circuit, radiating patch is represented by a parallel resonator ($R_a$-$L_a$-$C_a$). The fringing of the electric field from the radiating edges causes an increase in length of the patch. This additional length ($\Delta L$) is taken into account to determine the initial values of inductance ($L_a$) and capacitance ($C_a$). The value of $\Delta L$ is obtained using (10) [38]

$$\Delta L = 0.412 \frac{\varepsilon_{\text{eff}} + 0.3}{\varepsilon_{\text{eff}} - 0.258} \left( \frac{L_p}{\varepsilon_{\text{eff}}} + 0.264 \right)$$

The pair of T-shaped stubs is modeled as a parallel resonator with inductance $L_1$ and capacitance $C_1$. Two sections of transmission lines having lengths $l_1$ and $l_2$ and characteristics impedances $Z_{01}$ are used to represent the microstrip feed line. Here ($l_1 + l_2$) is the total length of the feed, which has a value

$$\rho_{\text{inksh}} = \frac{S_{21}}{1 - S_{11} (Z - 1) / (Z + 1)}$$
of 3.5 mm. The capacitance $C_a$ of the patch is determined from the parallel plate capacitor equation. While the $C_1$ is obtained using (11) for an open circuit TL theory

$$C_1 = \tan \beta l_o/2\pi f Z_{02}$$ (11)

Here $Z_{02} = 122.5 \Omega$ is the characteristics impedance of the stub for the width of 0.17 mm. The stub has an effective length ($l_o$) of 2.73 mm. Inductances $L_1$ and $L_2$ of the stub and patch are obtained using (12) by substituting corresponding length ($l$), width ($w$), metal thickness ($t$) in millimeters [39].

$$L(nH) = 0.21 \left[ \ln \left( \frac{l}{w+t} \right) + 1.193 + 0.2235 \left( \frac{w+t}{l} \right) \right]$$ (12)

In Fig. 7 (a) $C_i$ denotes the coupling capacitance between the open end of the microstrip feed and the square patch and $R_d$ signifies the radiation resistance of the antenna. The values of both parameters ($R_d$ and $C_i$) are found using the curve fitting approach. The equivalent circuit of the antenna is simulated using ADS and the reflection coefficient ($S_11$) is depicted in Fig. 7 (b). Full wave simulation of the proximity feed antenna without the SIW cavity is performed in HFSS. A good correlation is seen between the circuit simulated and HFSS simulated $S_11$. The antenna exhibits wideband impedance matching over the entire frequency sweep. Simulated $S_11$ of the proposed SIW cavity-loaded antenna is presented in Fig. 7 (b). This single antenna element shows a resonance dip at 28 GHz with a return loss of 38 dB. Within the 5G frequency range (27.5−28.35 GHz), the impedance matching is $> 20$ dB. Simulated 2D radiation patterns of the proposed single element antenna loaded with and without SIW cavity are illustrated in Fig. 8. The radiation patterns of the proposed antenna loaded with SIW cavity show significant improvement in FTBR in both principal planes- E and H, when compared with unloaded reference antenna. The antenna shows broadside radiation with FTBR of 22.3 dB which is more than 6.85 dB relative to the reference patch antenna. The antenna beamwidths become narrow after being loaded with the SIW cavity which also causes an improvement in antenna gain. The peak realized gain of the reference antenna and that of the SIW cavity-loaded antenna are 4.6 dBi and 6.7 dBi respectively.

![FIGURE 8. Simulated radiation patterns of the proximity feed antenna loaded with and without SIW cavity at 28 GHz in (a) E-plane (b) H-plane.](image)

**IV. MUTUAL COUPLING REDUCTION**

**A. H-PLANE COUPLED MIMO ANTENNA**

The wideband bandgap property of the proposed EBG as described in section II is utilized for mutual coupling reduction between mm-wave antennas in 5G MIMO, depicted in Fig. 9. The antenna elements are excited using proximity feed method with electromagnetic (EM) wave is guided through SIW based transition. The incorporation of SIW which is a planar form of rectangular waveguide ensures high isolation at the input ports and diminishes the radiation leakage losses from the circuit. Two antennas can transmit and receive the EM waves independently and make a 2-element MIMO antenna. The proposed MIMO antenna consists of two substrates, having a SIW feed network built on Taconic TLY-5 (height $h_1 = 0.51$ mm) and the antennas with EBG on FR-4 material (height $h_2 = 0.8$ mm). At input sections of the network grounded coplanar waveguide to SIW transition is designed to facilitate the measurement. Two 2.92 mm end launch connectors from Southwest are attached with the input ports. At the output of SIW, EM wave is directly coupled from the open end of microstrip line to the radiating patches. In the proposed MIMO both radiating patches are enclosed within the SIW cavity. This improves the radiation patterns of the antenna and increases the FTBR.

Fig. 10 shows the schematics of three metal layers of the proposed MIMO antenna. The first two layers are the DGS incorporated ground plane and SIW feeding, printed on two sides of Taconic TLY-5 substrate. The third metal layer is printed on a single side of FR-4 substrate. It consists of two patch antennas with symmetrically placed 2 $\times$ 3 EBG unit cells in the middle. Two antennas are H-plane coupled with the center to center spacing ($d$) between them is 6.2 mm or 0.58 $\lambda_0$, where $\lambda_0$ is the wavelength at 28 GHz.

**B. SIMULATION RESULTS**

Simulated return loss and isolation characteristics of the MIMO are depicted in Fig. 11. The reference MIMO
FIGURE 10. Layout of the proposed decoupled MIMO array (a) bottom ground plane with inset shows hair-pin DGS (b) middle layer consists of SIW feed (c) top layer comprises of EBG loaded antennas enclosed by SIW cavity. Parameters are $L = 31.1$, $W = 34.7$, $L_d = 21$, $W_d = 22$, $w_a = 6.2$, $l_c = 10.5$, $w_c = 15$, $L_p = 1.8$, $P_y = 8.3$, $C = 1.2$, $d_y = 1.3$, $d_x = 1.35$, $w_r = 0.18$, $g_r = 0.15$, $t_1 = 2$, $t_2 = 0.85$, $w_1 = 1.1$, $w_2 = 0.19$, $l_e = 5.8$, $w_e = 0.96$, $d = 6.2$.

FIGURE 11. Simulated (a) reflection and (b) transmission coefficients of the proposed MIMO antenna without and with decoupling elements.

FIGURE 12. Simulated radiation patterns of the MIMO antenna at 28 GHz without and with decoupling elements (a) E-plane (b) H-plane.

(without EBG-DGS) exhibits resonance at 28.04 GHz and shows impedance matching above 21.6 dB over 27.5–28.35 GHz, assigned for 5G application by Federal Communications Commission (FCC). Placing of EBG unit cells in the middle of two antennas reduces the coupling field. The maximum improvement in isolation is observed at 27.5 GHz (value of 23.4 dB) with an average enhancement in $|S_{21}|$ being 13.9 dB over the 5G range. It is observed in simulation that with a change in physical parameters ($S$ and $R_i$) of the proposed EBG, the frequency dip in $S_{21}$ shifts toward lower frequency away from the 5G operating range. Also the mutual coupling can not be reduced beyond a certain limit with only the EBG loading. Isolation between the antennas is further improved by placing an HP-DGS in the ground plane. DGS is symmetrically placed underneath the EBG as shown in Fig. 10 (a). The design parameters and the distance ($P_y$) of the DGS are optimized to achieve maximum isolation at 28 GHz. The proposed EBG-DGS based hybrid decoupling method shows an improvement in $|S_{21}|$ of 47.7 dB at 27.94 GHz relative to the reference array. The average in band isolation of the MIMO antenna with only EBG loading is 31.9 dB while that after loading with EBG-DGS is 39.2 dB. Fig 12 shows the simulated radiation pattern of the 5G MIMO antenna in two principal planes (E and H) at frequency 28 GHz. The loading of EBG and HP-DGS hardly causes any change in radiation patterns in both planes. Placing of decoupled MIMO antenna within SIW cavity results in high FTBR with simulated maximum value is 26.1 dB. Isolation improvement between the radiating patches can be explained physically from coupling
FIGURE 13. Surface current distribution on patch surface at 27.94 GHz (a) unloaded reference MIMO (b) EBG DGS loaded MIMO.

C. DIVERSITY PERFORMANCE OF THE MIMO

The parameters which are commonly used to describe the diversity of the MIMO are envelope correlation coefficient (ECC) and diversity gain (DG). Low correlation between the antenna elements is needed for the MIMO system, which can be analyzed using ECC. In practice, ECC should be < 0.5 to ensure high channel capacity or a good diversity performance. Two common methods of determining the ECC are based on scattering parameters and radiating fields. ECC ($\rho_e$) obtained using the S-parameters has less accuracy for antenna radiation efficiency < 90%. It is preferable to estimate the ECC from far-field radiation pattern using (13) [40]

$$\rho_e = \frac{\left| \iint \overrightarrow{F}_1(\theta, \varphi)^* \overrightarrow{F}_2(\theta, \varphi) d\Omega \right|^2}{\iint \left| \overrightarrow{F}_1(\theta, \varphi) \right|^2 d\Omega \times \iint \left| \overrightarrow{F}_2(\theta, \varphi) \right|^2 d\Omega}$$ (13)

Here $\overrightarrow{F}_1(\theta, \varphi)$ represents the radiation field pattern of the antenna. Fig. 15 shows ECC of the proposed MIMO antenna.
FIGURE 16. Photographs of SIW feed network (a) top view (b) bottom view (inset shows HP-DGS on backside ground plane) (c) photograph of EBG-DGS loaded MIMO antenna (inset shows the radiating patches with EBG unit cells).

FIGURE 17. Radiation pattern and gain measurement setup inside anechoic chamber (inset shows the fabricated prototype of EBG-DGS loaded MIMO antenna).

FIGURE 18. Simulated and measured (a) return loss and (b) isolation characteristics of the decoupled MIMO antenna.

FIGURE 19. Simulated reflection coefficient of the EBG-DGS loaded MIMO antenna at different air spacer heights.

V. EXPERIMENTAL RESULTS OF MIMO ANTENNA

A prototype of the proposed MIMO antenna is fabricated and experimentally tested to validate the simulation results. Fig. 16 shows the photograph of the fabricated MIMO loaded with EBG and HP-DGS. Two substrates are bonded together using a thin layer of epoxy resin. The total size of the MIMO antenna is $34.7 \times 31.1 \times 1.31 \text{ mm}^3$. Scattering parameters of the antenna are tested using Keysight N5224B Vector Network Analyzer (VNA). The radiation pattern and gain of the MIMO antenna are measured inside an anechoic chamber at IIT Palakkad using measurement setup shown in Fig. 17. Double ridged broadband horn antenna operating between 10 to 40 GHz is used as a transmitting antenna whose distance from the AUT (antenna under test) is 30 cm to satisfy farfield condition. Fig. 18 depicts the measured reflection and transmission coefficients of the proposed MIMO, with results are compared with simulation. The little discrepancy occurs due to fabrication imperfection and the presence of a thin bonding layer in between the stacked dielectric substrates, resulting into change in effective dielectric constant. Measured $|S_{11}|$ and $|S_{22}|$ are > 17.4 dB within the 5G range. A dip in
measured $|S_{21}|$ occurs at 27.76 GHz which is shifted to lower frequency as compared to the simulated response. Isolation characteristic closely follows simulated $|S_{21}|$. Maximum values of $|S_{21}|$ are 71.9 dB in measurement and 65.4 dB in simulation. Within the mm-wave 5G operating range in band isolation is $>32.7$ dB. The tolerance of reflection coefficient on the presence of thin layer of air having height ($h_{air}$) between two substrates can be further analyzed in simulation. Fig. 19 shows that increase in air thickness causes a shift in resonance frequency towards lower frequency. The above result is found consistent with the measured reflection coefficient of the MIMO antenna.

While measuring the radiation patterns one of the antennas is matched terminated with 50 $\Omega$ broadband load. The E and H plane patterns at three frequencies 27.75, 28, and 28.25 GHz inside 5G range are measured and compared with the simulated results, as shown in Fig. 20. Measured radiation pattern is in good accordance with simulated result. Patterns exhibit a boreside radiation in both principal planes. FTBR has a maximum measured value of 19.8 dB, which is little lower than the simulated value. The gain of the decoupled MIMO antenna is also measured using gain comparison method. In this method power received by the AUT is compared with that received by standard gain horn antenna. Fig. 21 presents the simulated and measured gain of the MIMO antenna within 5G operating range. At frequency 28 GHz, the measured peak gain is obtained as 9 dBi against the simulated value of 9.3 dBi. At high frequency measured gain decreases from the simulated value. This can be attributed to an increase in losses due to radiation leakages and higher coupling between the antenna elements. Fig. 22 shows the simulated and measured total efficiency of the antenna. Wheeler cap method is realized to experimentally determine the efficiency of the antenna. The antenna is placed inside a cylindrical cavity made of thin layer of aluminum foil. The reflection coefficients of the antenna are measured inside the cavity and also in the outside free space. The two reflection coefficients are denoted as $S^{WC}_{11}$ and $S^{FS}_{11}$ respectively from which the efficiency ($\eta$) is found using the relation (14) [41]

$$\eta = 1 - \frac{(1 - S^{FS}_{11})(1 + S^{WC}_{11})}{(1 + S^{FS}_{11})(1 - S^{WC}_{11})}$$

(14)

At 28 GHz the simulated and measured values of $\eta$ are close to 82 % while that over the 5G operating band is $>73.4$ %. In the MIMO system, channel capacity signifies the maximum data rate for reliable transmission of information between the transmitter and receiver. Channel capacity usually increases with an increase in number of antennas. The mutual coupling between the antennas causes losses in the channel capacity. The CCL is found from the correlation matrix ($\psi^k$) using (15) [42]

$$CCL = - \log_2 \det \left( \psi^k \right)$$

(15)

Elements of coupling matrix $\psi^k = \begin{bmatrix} \rho_{11} & \rho_{12} \\ \rho_{21} & \rho_{22} \end{bmatrix}$ is related to the S-parameters using (16) [42]

$$\rho_{ii} = 1 - (|S_{ii}|^2 + |S_{ji}|^2), \quad \rho_{ij} = - \left( S_{ii}^* S_{ij} + S_{jj}^* S_{ji} \right)$$

(16)

for $i, j = 1 \text{ or } 2$. Acceptable value of CCL is $<0.4$ bits/Hz/sec for the MIMO system. The simulated and measured CCL of the proposed MIMO antenna are shown in Fig. 23. The values

| Ref Year | [5] 2017 | [8] 2016 | [13] 2016 | [14] 2015 | [22] 2017 | [24] 2016 | [25] 2020 | [30] 2019 | [31] 2018 | [32] 2017 | [33] 2019 | [34] 2017 | This work |
|----------|------------|------------|------------|------------|------------|------------|------------|------------|------------|------------|------------|------------|------------|
| Method   | Coupled line resonator | Polarization conversion isolator | Slot combined CSRR | 1D EBG + SRR | EBG + DGS | Fractal DGS | Absorber | Hybrid isolator | CSRR | Meta-surface | Metal strip | FSS wall + DGS | EBG + DGS |
| Peak Isolation (dB) | 40 | 32.8 | 52 | 53.7 | 35 | 60 | 43.71 | 49 | 55 | 49.8 | 24 | 36 | 71.9 |
| In Band Isolation (dB) | 20 | 25 | 35 | 14 | 22 | 32 | 23 | 29 | 32 | 35 | 22 | 27 | 32.7 |
| Improvement in $|S_{11}|$ (dB) | 26.2 | 22.3 | 27 | 42.1 | 22 | 37 | 12.41 | 25 | 31.8 | 20.12 | 12 | 21.5 | 47.7 |
| Gain (dBi) | 6.25 | $<$ 5 | 3.59 | 2.57 | $<$ 5 | 5 | 7.74 | 8.1 | NA | 17.9 | 8 | NA | 9 |
| Efficiency (%) | 87 | 60 | NA | 82 | 75 | 96 | 68.03 | 69 | NA | 85 | NA | 90 % | 81.9 |
| FTBR (dB) | 16 | 13 | 20 | 9.3 | 10 | 9 | 15.2 | 11 | NA | 20 | $<$ 10 | 10 | 19.8 |
| ECC       | NA | NA | NA | 0.002 | 0.002 | $-$175 dB | 0.05 | NA | NA | NA | 0.013 | $5 \times 10^{-6}$ | 0.00015 |
| Element Spacing ($\lambda_0$) | 0.29 | 0.39 | 0.36 | 0.19 | 0.36 | 0.38 | 0.55 | 0.5 | 0.67 | 1.96 | 0.82 | 0.5 | 0.58 |
FIGURE 20. Simulated and measured E and H plane radiation patterns of the EBG DGS loaded MIMO antenna at (a) 27.75 GHz (b) 28 GHz (c) 28.25 GHz.

FIGURE 21. Simulated and measured gain of the EBG DGS loaded MIMO antenna.

FIGURE 22. Simulated and measured efficiency of the MIMO antenna loaded with EBG and DGS for coupling reduction.

FIGURE 23. Simulated and measured CCL of the EBG DGS loaded MIMO antenna.

VI. SIW CAVITY BACKED 4-ELEMENT LINEAR ARRAY

A. DESIGN OF LINEAR ARRAY

A SIW cavity-backed 4-element linear array is designed to illustrate the effect of the hybrid decoupling method on the array performance. The array is fed with a 4-way SIW power divider. Radiating patches are proximity coupled with output ports of the power divider. The front and back geometries of the proposed linear array are shown in Fig. 24. The proposed EBG and HP-DGS based hybrid decoupling elements are placed between the array elements to reduce the coupling between the adjacent antennas. Low mutual coupling between the antennas improves the overall impedance matching of the linear array. At 28 GHz the output signal amplitude at the four ports of the SIW power divider is 6.75 dB. The additional 0.75 dB loss at each output port is attributed to the effects of dielectric loss and reflection loss from the two stages T-junction dividers due to impedance mismatch. The work has the advantage of higher isolation improvement among previously reported works. The value of ECC is also found lower than all reported works except [34]. Compared to mm-wave MIMO in [30]–[34], present work shows major improvement in $|S_{21}|$ with a comparable antenna spacing. However, the gain in [32] is higher than the proposed MIMO antenna. The antenna element used in [32] is an antipodal Vivaldi with Fermi tapered slot that incurs more design complexity and has a larger footprint.
The total power radiated by 4-antennas are added in phase in the broadside direction and improve the gain of the array in H-plane. As previous, the array elements are enclosed by a SIW cavity to improve the FTBR at the resonance. The lateral dimension of the linear array is $43 \times 40 \text{ mm}^2$ or $4.01 \times 3.73 \lambda_0^2$, where $\lambda_0$ is the wavelength at 28 GHz. A prototype of the array is also fabricated and experimentally tested. SIW power divider is made on Taconic TLY-5 substrate, whose backside is a HP-DGS integrated ground plane. The four radiating patches with three groups of $2 \times 3$ EBG unit cells are fabricated on FR-4 substrate. The substrate heights of Taconic TLY-5 and FR-4 are 0.51 mm and 0.8 mm respectively. Silver paste is used to fill the via holes in the structures. Photographs of the 4-way power divider and the prototype of the EBG-DGS loaded 4-element linear array are presented in Fig. 25.

**FIGURE 24.** Proposed SIW cavity backed 4-element linear array loaded with EBG-DGS (a) front view (b) back view. $L_a = 40$, $W_a = 43$, $l_d = 13.2$, $w_d = 36$, $d_a = 7.5$.

**FIGURE 25.** Photographs of 4 way SIW power divider (a) top view (b) bottom view (inset shows the HP-DGS) (c) Photograph of the fabricated 4-element linear array (inset shows EBG unit cells loaded radiating patches).

**FIGURE 26.** Simulated and measured reflection coefficients of the EBG-DGS loaded and unloaded 4-element linear array.

### B. SIMULATION AND MEASUREMENT RESULTS

The simulated and measured reflection coefficients of the linear array are depicted in Fig. 26. The resonance dip of the array occurs at 28 GHz in simulation and 27.85 GHz in measurement with the return losses are 36.3 and 33.8 dB respectively. The $S_{11}$ of the array is compared to the EBG-DGS unloaded reference array and the simulation result shows an improvement of 13.1 dB in the return loss magnitude. A small shift of resonance frequency in the measured $S_{11}$ is present that may arise due to fabrication imperfection.
Also, change in relative permittivity due to the use of thin layer of epoxy resin as bonding film affects the measured response of the array. The measured 10 dB return loss BW of the array covers the frequency range 27.58 to 28.17 GHz. The simulated and measured gain variations of the proposed 4-element linear array are shown in Fig. 27 (a). Peak realized gain of the array at 28 GHz is 12.3 dBi in simulation and 13.3 dBi in measurement. The H-plane radiation pattern of the array at the resonance is depicted in Fig. 27 (b). The pattern shows a low sidelobe level (SLL) of 13.1 dB and FTBR of 19.97 dB in the measurement. FTBR little degrades as compared to simulated values of 27.5 dB, which may occur due to errors in the measurement. The 3 dB beamwidth of the array in H-plane is 16.2°. Fig. 28 shows the simulated and measured efficiencies of the 4-element linear array. The measured and simulated values of the efficiency are 75.9 % and 81.7 % at the operating frequency of the linear array.

VII. CONCLUSION

A novel uniplanar miniaturized EBG unit cell is designed. The proposed EBG exhibits a wideband bandgap of 20 % which has promising application for mm-wave circuit designs. This paper explores the application of EBG as a decoupling element for mutual coupling reduction in MIMO antenna. A newly proposed hair-pin shaped DGS is also included in the ground plane to further improve the isolation. The simulation and measurement results show high value of isolation over the 5G range that has potential to improve the channel capacity or beam scanning angle of the phased array. The MIMO shows the isolation of 71.9 dB between its elements with ECC value of 0.00015. The proposed MIMO antenna has the advantage of good unidirectional patterns in two principal planes with a measured peak gain of 9 dBi. The hybrid decoupling technique can be further extended for massive MIMO configuration with proper adjustment in the feed. A SIW cavity-backed 4-element linear array is also designed with EBG-DGS based decoupling elements are placed in between the antenna elements. Good radiation characteristics with low SLL, narrow beamwidth, high gain are observed for the proposed array. Indigenous designs of SIW cavity and feeding network are presented to obtain high FTBR and improved radiation patterns of the MIMO antenna and linear array, which will find application in future 5G communication at mm wave band.

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SUomal Dey (Senior Member, IEEE) received the B.Tech. degree in electronics and communication engineering from the West Bengal University of Technology, Kolkata, India, in 2006, the M.Tech. degree in mechatronics engineering from the Indian Institute of Engineering Science and Technology (IIEST), Shibpur, India, the M.Tech. dissertation (one year) from Central Electronics Engineering Research Institute (CEERI-CSIR), Pilani, India, in 2009, and the Ph.D. degree from the Centre for Applied Research in Electronics (CARE), Indian Institute of Technology Delhi, in July 2015.

From August 2015 to July 2016, he was working as a Project Scientist with Industrial Research and Development (IRD) Centre, IIT Delhi. He also worked on a collaborative research project supported by Synergy Microwave Corporation, NJ, USA, during the same period. From August 2016 to June 2018, he was working with the Radio Frequency Microsystem Lab (RFML), National Tsing Hua University, Taiwan, as a Postdoctoral Research Fellow. Since June 2018, he has been working as an Assistant Professor with the Department of Electrical Engineering, Indian Institute of Technology Palakkad, Kerala. He was a recipient of the Postgraduate Student Award from the Institute of Smart Structure and System, Bengaluru, in 2012; the Best Industry Relevant Ph.D. Thesis Award from the Foundation for Innovation in Technology Transfer, IIT Delhi, in 2016; the Postdoctoral Fellow Scholarships from the Ministry of Science and Technology (MOST), Taiwan, in 2016 and 2017; the Early Career Research Award (ECRA) from the Science and Engineering Research Board (SERB), Government of India, in 2019; and the Smt. Ranjana Pal Memorial Award from the Institution of Electronics and Technology (IETE), in 2021. He has authored/coauthored more than 70 research articles, one state-of-the art book, two book chapters, and holds eight patents. His research interests include RF MEMS, microwave, sub-mm wave metamaterial structures, and microwave integrated circuits, including antennas.

Shiban K. Kou (Life Fellow, IEEE) received the B.E. degree in electrical engineering from Regional Engineering College, Srinagar, India, in 1977, and the M.Tech. and Ph.D. degrees in microwave engineering from the Indian Institute of Technology Delhi, New Delhi, India, in 1979 and 1983, respectively.

He has been an Emeritus Professor with the Indian Institute of Technology Delhi, since 2019, and the Mentor Deputy Director (Strategy & Planning, International affairs) of IIT Jammu, Jammu and Kashmir, India, since 2018. He served as the Deputy Director (Strategy and Planning) of IIT Delhi, from 2012 to 2016. He also served as the Chairperson of Astra Microwave Products Ltd., Hyderabad, from 2009 to 2019, and Dr. R. P. Shenoy Astra Microwave Chair Professor at IIT Delhi, from 2014 to 2019. His research interests include RF MEMS, high frequency wireless communication, microwave engineering, microwave passive and active circuits, device modeling, millimeter and sub-millimeter wave IC design, body area networks, flexible and wearable antennas, medical applications of sub-terahertz waves, and reconfigurable microwave circuits, including miniaturized antennas. He has successfully completed 38 major sponsored projects, 52 consultancy projects, and 61 technology development projects. He has authored/coauthored 550 research papers, 17 state-of-the art books, five book chapters, and two e-books. He holds 26 patents, six copyrights, and one trademark. He has guided 27 Ph.D. theses and more than 120 master’s theses. He is a fellow of INAE and IETE. He is the Chief Editor of IETE Journal of Research, an Associate Editor of the International Journal of Microwave and Wireless Technologies (Cambridge University Press). He served as a Distinguished Microwave Lecturer of IEEE MTT-S for the period 2012–2014. Prior to this, he served as a Speaker Bureau Lecturer for IEEE MTT-S. He also served as an AdCom Member of the IEEE MTT-S, from 2010 to 2018, and is currently a member of the Awards, Nomination and Appointments, MGA, M&S, and Education committees of the IEEE MTT-S. He was a recipient of numerous awards, including the IEEE MTT Society Distinguished Educator Award (2014); the Teaching Excellence Award (2012) from IIT Delhi; the Indian National Science Academy (INSA) Young Scientist Award (1986); the Top Invention Award (1991) of the National Research Development Council for his contributions to the indigenous development of ferrite phase shifter technology; the VASVIK Award (1994) for the development of Ka-band components and phase shifters; the Ram Lal Wadhwa Gold Medal (1995) from the Institution of Electronics and Communication Engineers (IETE); the Academic Excellence Award (1998) from Indian Government for his pioneering contributions to phase control modules for Rajendra Radar; the Shri Om Prakash Bhasin Award (2009) in the field of electronics and information technology; the VASVIK Award (2012) for the contributions made to the area of Information, Communication Technology (ICT); and M. N. Saha Memorial Award (2013) from IETE.