60 GHz 3D Passive Electromagnetic Deflector for Wide Angular Coverage

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Abstract—A 3D passive electromagnetic deflector configuration, based on the concept of spatial-feeding, is proposed that modifies the incident electromagnetic energy from the steerable planar antenna array (source) to achieve wide angular coverage. A generalized analytical formulation of the parametric deflector configuration is carried out. The analytical modelling has been validated by 3D simulations and measurements. The analytical results for the deflector configuration indicate that a wide angular coverage is possible at the cost of reduced gain. An improved 60 GHz circularly-polarized (CP) deflector element has been designed, which provides phase shift with element rotation upon transmission of circular polarization. The simulation results show that the proposed CP element operates successfully, in terms of realization of phase-shift range from 0° to 360° and good transmission properties (transmission loss < 1dB) for the desired frequency band of 57–64 GHz.

1. INTRODUCTION

The unlicensed frequency band around 60 GHz seems to be very promising in the realization of the goal of broadband wireless communication over short distances. The usage models (wireless sync, wireless high-definition (HD) display, wireless peripheral interconnection and wireless internet) are described in the IEEE standard 802.11/ad (WiGig) for 60 GHz applications, having line-of-sight (LOS) or non-line-of-sight (NLOS) operation [1]. As regards the first feasible commercial products, the short-range LOS applications for peer-to-peer communication (kiosk download and wireless docking) are of prime interest [2]. The link-budget requirement for these application scenarios could be met with the use of a small antenna array [3]. However, to utilize the full potential of the 60 GHz frequency band for an interesting subset of the usage models (like wireless HD home-theater scenario), the antenna technology is a hurdle and it is desirable to have antennas with wide angular coverage.

To-date, a very few 60 GHz wide angular coverage antennas are reported in literature. The 60 GHz angle-diversity antenna array, based on seven dielectric rod antennas, offers an advantage of a simple design [4]. The switched-beam antenna array provides the scan-range of ±50° with a gain variation from 6.8–11 dBi. The wide-angular coverage of ±45° with maximum gain and gain scan-loss of 21.7 dBi and 1.1 dB, respectively, have been shown using a mechanically-steerable lens antenna [5]. The simple mechanically-steerable lens antenna cannot offer electronic beam-steering. The quasi omnidirectional antenna behaviour is achieved using a 60 GHz shaped reflector [6] and shaped-beam lenses [7]. A conceptual design of 60 GHz multifaceted active antenna array (MF AAA) is mentioned in [8], comprising multiple levels of 1 × 8 subarrays of antenna elements with a central RFIC scheme. The simulation results show that the broadside gain of 13 dBi can be achieved by the use of multiple subarrays. The 60 GHz RF interconnection losses from the RFIC to each antenna element are not addressed and the significant losses are expected at 60 GHz. MF AAA, in principle, can provide a hemispherical coverage but the limitation is primarily dictated by the RF interconnection from the central RFIC unit to the active antenna array on each facet.

The antennas for 60 GHz applications, like wireless HD home-theater, require electronic steering and minimum associated losses. The antenna gain should also be high enough (typically > 10 dBi) [3], [9] to satisfy the minimum link budget requirement with almost hemispherical scan coverage. Moreover, for maximum utilization of the allowed effective isotropically radiated power (EIRP) over the whole scan region, the antenna gain envelope should be as constant as possible [10]. In view of the stated figure-of-merit, the reviewed 60 GHz literature for wide angular coverage is summarized in Table I.

<table>
<thead>
<tr>
<th>Principle</th>
<th>Ref.</th>
<th>Electronic steering</th>
<th>Antenna gain requirement</th>
<th>RF interconnect losses</th>
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<tr>
<td>MF AAA</td>
<td>[8]</td>
<td>+</td>
<td>+</td>
<td>-</td>
</tr>
<tr>
<td>Shaped-beam lens</td>
<td>[6], [7]</td>
<td>-</td>
<td>-</td>
<td>+</td>
</tr>
<tr>
<td>Angle-diversity</td>
<td>[4]</td>
<td>+</td>
<td>-</td>
<td>+</td>
</tr>
<tr>
<td>Steerable lens</td>
<td>[5]</td>
<td>-</td>
<td>-</td>
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</table>

* strong and weak points are labelled by + and -, respectively.

A 3D passive electromagnetic deflector configuration is proposed in Section II, as an alternative to the listed 60 GHz wide angular coverage antenna solutions. A generalized analytical formulation of the parametric 3D deflector configuration is described in Section III. Section IV highlights the validation of the analytical model for the single deflector facet setup with 3D simulations and measurements. After validation, the multifaceted deflector configuration is analyzed with the analytical model and compared with a spherical deflector for wide angular coverage in Section V. The performance
evaluation of the three different design schemes for the 60 GHz deflector element is carried out in Section VI and the design of an improved CP deflector element is presented. Section VII presents the conclusions.

II. PROPOSED DEFLECTOR CONFIGURATION

The beam of planar antenna arrays cannot usually be steered more than ±50° due to increased mismatches at large scan angles [11], [12], [13]. The gain envelope (envelope of the peak gain of the scanned antenna beams) vs scan coverage profile of a typical planar antenna array source and the desired wide angular coverage profile are shown in Fig. 1a(i). A 3D passive electromagnetic deflector, based on the concept of spatial-feeding [14], [15], is proposed for scan-range extension of a steerable planar antenna array (source); the artist impression is shown in Fig. 1b. In principle, the deflector based configuration modifies the incident electromagnetic energy from the steerable planar antenna array (source). By introducing passive deflector elements with fixed phase-shifts on the spherical deflector surface, a beam can be produced having a modified scan-angle \(\theta_S\) relative to the scan-angle of the emitted beam from the steerable planar antenna array source \(\theta_P\) (Fig. 1a(ii)). If \(\theta_P\) corresponds to the position of each deflector element then the fixed phase-shifts can be defined based on the ratio of \(\theta_S\) and \(\theta_P\), termed as scan amplification factor \(K_{\text{scan}} = \frac{\theta_S}{\theta_P}\).

\(K_{\text{scan}} = 1\) represents the case as if no deflecting surface is present on top of the source. The increased gain with a reduction in scan-range of the source can be achieved by using \(K_{\text{scan}} < 1\). The scan-range extension of the source for wide-angular coverage can be met with \(K_{\text{scan}} > 1\) at the cost of reduced gain, while satisfying the minimum antenna-gain requirement. The multifaceted deflector configurations serve as a counterpart of the spherical deflector design. It is desirable to use a CP deflector element with a CP planar antenna array source to minimize the polarization mismatch for all the azimuthal angles.

III. MODELLING OF 3D DEFLECTOR

The suggested deflector configuration is very large in terms of wavelength, so it is unfeasible to analyze it with commercially-available electromagnetic tools due to the computationally-intensive nature of the problem. A generalized analytical formulation of the parametric deflector configuration is hence needed, which is the topic of this section. The spherical deflector has been considered first, followed by the analysis of multifaceted deflector configuration for the realization in planar technology.

A. Spherical deflector

The general modelling of the spherical deflector including the geometrical aspects, analysis of the configuration and an approach of phase-shift determination for wide angular coverage is discussed in detail in [16].
Flow-chart of geometrical modelling of a 3D faceted deflector

**Step - I**
- a) Make facets from defined vertices of the desired geometry in GCS
- b) Determine direction cosines of the normal to the plane facet

**Step - II**
- a) Determine origin of the local-axis for each facet
- b) Determine direction cosines of the local-axis

**Step - III**
Coordinate transformation from GCS to LCS

**Step - IV**
Rectangular grid of deflector elements within the boundary of facet

**Step - V**
Coordinate transformation from LCS to GCS

Fig. 2. Flow-chart of geometrical modelling of a 3D faceted deflector configuration; GCS and LCS stand for global coordinate system and local coordinate system, respectively.

which are complex-valued functions of spatial coordinates only (i.e., time dependence is not shown). The conversion from phasor quantities to all real-time varying quantities can be accomplished by multiplying the phasor by \( e^{j\omega t} \) and taking the real part [18], where \( j \) is the imaginary unit, \( \omega \) is the angular frequency, and \( t \) is the time. The scalar far-field of the reference elementary antenna of a source antenna array \( E_s(r, \theta, \phi) \) at the far-field distance \( r \) in the upper half-space \((z > 0)\) can be represented by \( E_s(r, \theta, \phi) = \frac{1}{r} \sum_{m=1}^{M} I_m e^{jkr}, \) where \( I_m \) is the total radiated power of the source antenna array in the upper half-space which is given by

\[
P_{source} = \frac{1}{2\eta_0} \int_0^{2\pi} \int_0^\pi |E_{source}(r, \theta, \phi)|^2 r^2 \sin \theta d\theta d\phi. \tag{3}\]

The gain of each deflector element \( G_F(\theta_F, \phi_F) \) on the reception and the transmission sides is modelled as \( G_F(\theta_F, \phi_F) = \frac{1}{\lambda_0^2} A_{cell} \cos \theta_F, \) where \( 0 \leq \theta_F \leq \frac{\pi}{2}, \) and \( 0 \leq \phi_F < 2\pi \) is the design wavelength. The directivity of each deflector element is related to its gain by \( G_F(\theta_F, \phi_F) = D_F(\theta_F, \phi_F)e_{rad}, \) where \( e_{rad} \) is the radiation efficiency. Since every element \( m \) on a given facet \( F \) sees the far-field observation point at the same angle, element’s subscript \( m \) is replaced by facet’s subscript \( F. \)

The available power \( P_{cm} \) on the reception side of each deflector element in the \( xyz \)-space can be written in terms of the power flow density \( S(R_m, \alpha_m, \beta_m) \) of the incident plane wave towards the direction of the deflector element \( m \) and the effective area of the deflector element \( A_e. \) The subscript \( m \) is re-introduced because \( P_{cm} \) depends on \( m. \) The polarization mismatch factor has not been considered, hence,

\[
P_{cm} = S(R_m, \alpha_m, \beta_m) A_e(\alpha_m, \beta_m) \cos \xi_m, \tag{4}\]

\[
S(R_m, \alpha_m, \beta_m) = \frac{|E_{source}(R_m, \alpha_m, \beta_m)|^2}{2\eta_0}, \tag{5}\]

where \( \xi_m \) is the angle between the radial unit-vector of the incident power flow density and the unit-vector normal to the deflecting element on the facet \( F \) and is determined using \( \xi_m = \arccos[\sin \alpha_F \sin \alpha_m \cos(\beta_m - \beta_F) + \cos \alpha_F \cos \alpha_m]. \)

The effective area of each deflector element is

\[
A_e = \frac{\lambda_0^2}{4\pi} G_F(\theta_F, \phi_F). \tag{6}\]

The power of each deflector element on the transmission side \( P_{tm} \) is related to the transmission coefficient \( |S_{21}^m| \) as \( P_{tm} = |S_{21}^m|^2. \) The transmission coefficient is a function of angle of incidence and can be determined from 3D EM simulations or measurements for a specific deflector element. The power of each deflector element on the transmission side \( P_{tm} \) is taken to be \( P_{cm} \) for the lossless case. The far-field electric field of the faceted deflector \( E_{def}(r, \theta, \phi) \), comprising \( N_F \) facets with \( M \) deflector elements on each facet, and the directivity \( D_{def}(\theta, \phi) \) of the deflector configuration in the upper half-space \((0 \leq \theta \leq \frac{\pi}{2}\) and \(0 \leq \phi < 2\pi)\) can be calculated as

\[
E_{def}(r, \theta, \phi) = \exp(-jkr) \sum_{F=1}^{N_F} \sum_{m=1}^{M} E_m(r, \theta, \phi),
\]

\[
D_{def}(\theta, \phi) = \frac{|E_{def}(r, \theta, \phi)|^2}{2\eta_0}, \tag{7}\]
The phase distribution profile (from the origin $O$) of the spherical deflector to achieve wide angular coverage is shown in Fig. 3b. The total phase-shift at each element position $\alpha_m$, including the phase due to spatial-feed path-length $R$ from the center of the source $O$ to each element $n$ and the phase-shift provided by each element for wide angular coverage, can be formulated as $-k(R+l_m)$, where $k = \frac{\lambda}{2\pi}$ is the free-space phase constant and $R+l_m$ is the total path-length. As mentioned in [16], for wide angular coverage $l_m = \frac{R}{K_{\text{scan}}}[1-\cos(K_{\text{scan}}\alpha_m-\alpha_n)]$, with $K_{\text{scan}} = 1.5$.

- A heuristic approach is used to determine the phase distribution of the multifaceted deflector configuration from the spherical deflector’s phase formulation. The multifaceted deflector configuration is inscribed inside the sphere as shown in Fig. 3c. Intuitively, the phase distribution for the multifaceted deflector configuration should follow the spherical phase-profile in a piece-wise way according to the multifaceted deflector geometry, as illustrated in Fig. 3c. The multifaceted deflector geometry and the phase-profile tend to converge to the spherical case, as the number of faces are increased. The transformation from the spherical case is performed point-wise at the position of each deflector element $(R_m,\alpha_m,\beta_m)$, resulting in the total phase-shift for each element as $-k(R_m+l_m)$. The phase-shift of each deflector element $\gamma_m = -k l_m$ on the multifaceted deflector configuration for wide angular coverage with $K_{\text{scan}} = 1.5$ can be formulated as

$$\gamma_m = \frac{-k l_m}{K_{\text{scan}} - 1}[1-\cos(K_{\text{scan}}\alpha_m-\alpha_n)]. \quad (9)$$

There are some bounds, however, for the heuristic approach which are typically dictated by the deflector element size and the average radius of the multifaceted deflector configuration (radius of the circumscribed sphere). The phase-deviation is chosen as a figure-of-merit, defined as the difference between the phase-profile of the spherical deflector and the multifaceted counterpart, and is evaluated for a variable average radius and grid-spacing of the multifaceted deflector configuration. The results for the 2D analysis of the case-example (Fig. 3c) are presented in Fig. 4. The general remarks regarding the heuristic approach of phase determination are as under:

- The defined phase-deviation is reduced with an increase in number of faces for the multifaceted deflector configuration.
- As the average radius of the multifaceted deflector is reduced, the phase-deviation becomes smaller. However, the number of available phase-states with element size equal to or larger than $\frac{\lambda}{2\pi}$ results in a phase discontinuity which can be overcome by reducing the element dimensions to $\frac{\lambda}{2\pi}$ or less.
- The phase-resolution is improved with an increase in average radius of the deflector, as more deflector elements become available. The increased radius, however, gives a larger phase-deviation resulting in high-frequency ripples on the pattern. The ripples can also be observed with $\frac{\lambda}{2\pi}$ or less element size, although the fluctuation is less pronounced as with element size of $\frac{\lambda}{2}\pi$.

The heuristic approach of phase determination for the deflector configuration considerably reduces the design space and can be optimized further using numerical methods. The performance of multifaceted deflector geometry, however, is evaluated using the presented heuristic approach only.

IV. VALIDATION OF THE ANALYTICAL MODEL

It is important to validate the analytical model to have confidence in the results. This can be done by considering a deflector set-up which is within the computational capabilities of the commercially-available 3D EM tools and is not restrictive for measurement purposes. The analytical modelling, presented in the last section has been validated by considering a simple set-up, comprising a 60 GHz hexagonal 6-element antenna array and the 60 GHz 0° passive electromagnetic deflector, reported in [20] and [15], respectively. A comparison is made with CST MWS simulations and measurements to validate the analytical model. Although the simulation tools, for instance CST MWS, take into account most of the electromagnetic effects, it is computationally very intensive and CST MWS requires more than 13 million mesh cells, for the said simple set-up.

Since the analytical model does not include diffraction from the deflector’s facet edges, a fair comparison with the CST simulated model is only possible, when the deflector facet is surrounded by an RF absorber or an infinite metal
Fig. 3. (a) Coordinate system for far-field radiation pattern determination of 3D multifaceted deflector configuration, (b-c) Heuristic approach of phase-shift determination using the transformation from spherical deflector (solid) with phase-profile (dashed) to multifaceted deflector (solid-black) with phase-profile (solid-gray).

Fig. 4. Bounds for the heuristic approach of phase determination; the profiles of the defined phase-deviation vs average radius of the multifaceted deflector configuration (7λ₀ and 14λ₀) with deflector element unit-cell size of \( \frac{1}{4} \) (dashed) and \( \frac{1}{2} \) (solid), for the case-example (Fig. 3c) are shown.

Fig. 5. (a) Comparison of transmission coefficient of 0° deflector for 0° angle of incidence; measurement (solid), simulation [aligned] (dashed-dotted), simulation [misaligned] (dashed) (b) Simulation of effect of angle of incidence on 0° deflector’s transmission coefficient; -15° (dashed), 0° (solid), +15° (dashed-dotted).

highlighted in Fig. 6a. In the analytical model, the pattern of the deflector element, is approximated by \( \frac{1}{4\pi} A_{cell} \cos \theta \) and the radiation pattern of a 6-element hexagonal source antenna array generated by CST MWS simulations has been used in the model.

CST MWS simulations use the total radiated power (upper and lower half-spaces) of the whole configuration as a normalization factor for the calculation of the gain. Since MATLAB formulation computes the gain of the whole deflector in the upper half-space \( (0 \leq \theta \leq \frac{\pi}{2} \text{ and } 0 \leq \phi < 2\pi) \) with total radiated power in the upper half-space as a normalization
Fig. 6. (a) CST model of a 6-element source antenna array and a 0° (9 × 9) broadside deflector facet for the 60 GHz band. (b) Measurement set-up: source antenna array seen from the top of the hollow window in the RF absorber inside a metal box (top-left), illuminated deflector area towards the source antenna array, showing (9 × 9) elements (top-right), 60 GHz anechoic chamber set-up for validation of the analytical model (bottom).

factor, the same has been used for computing the realized gain from the CST MWS results. In the analytical model, the transmission coefficient of each deflector element is determined from the results presented in Fig. 5 [17]. The angle of incidence dependence is incorporated for each deflector element as per its angle in the θ-domain, using extrapolation and interpolation.

The measurement set-up to validate the analytical model is shown in Fig. 6b, which is consistent with the CST simulated model. As is clear from Fig. 6, in CST MWS the ground plane of the deflector plate is extended to infinity, whereas, in the measurement set-up, the metal is backed by RF absorber and the area apart from the deflector plate’s illuminated window is covered by the RF absorber. The latter provides better immunity in terms of diffraction from the illuminated window. A 0° deflector plate comprising 9 × 9 elements is excited by a 6-element hexagonal source antenna array from the broadside direction. The broadside deflector facet is placed in the far-field of the source antenna array resulting in the height between the center of the broadside deflector facet and the source antenna array as 11.4λ₀, which is set by the limitations of the measurement set-up. Considering the illuminated deflector area of 9 × 9 elements and the height, the effective angular region of interest for radiation patterns of the deflector configuration is ±15°.

The CST simulated and measured patterns of a 6-element hexagonal source antenna array are depicted in Fig. 7a. Figure 7b shows the radiation patterns, when a 0° deflector plate comprising 9 × 9 elements is excited by the source antenna array from the broadside direction. A close agreement among the results of the analytical model, CST simulation and the measurement can be seen. Since the CST simulation set-up only uses the metal plate for surrounding the deflector without the backing by RF absorber, the relatively high ‘side-ripples’ can be observed in Fig. 7b. The difference (within 1.5 dB) in the maximum value of the deflector’s realized gain is attributed to less realized gain of the measured source array (Fig. 7a), consideration of transmission coefficient of aligned deflector elements (without fabrication misalignment) in the analytical model and CST simulation set-up and mathematical approximation of the radiation pattern of the deflector element in the analytical model. A slight asymmetry in the measured radiation pattern can be seen because the illuminated deflector area is not exactly in the middle of the 6-element source antenna array.

The close agreement among the analytical model, CST simulation and measurement validates the presented analytical model and gives confidence in the results of the extended analytical model for multifaceted deflector configurations.

V. SPHERICAL VS FACETED DEFLECTOR - RESULTS AND DISCUSSION

As a case study, the deflector configurations shown in Fig. 8 are analyzed. The simulation results of lossless spherical deflector and half truncated icosahedral deflector (HTID) configurations from the analytical model of Section III are highlighted in this section. The directivity patterns for source-HTID are compared with source-spherical deflector geometry for the same simulation set-up. As for the simulation set-up of the deflector configurations, a 19-element concentric circular array (CCA), shown in Fig. 8a, is chosen as a source antenna array. A CCA is chosen
as a source antenna array due to its symmetry in the whole azimuth plane. Moreover, concentric circular arrangement (2-rings) results in 19-elements, which can also satisfy the minimum antenna gain requirement. The deflector are placed at a distance of $7\lambda_0$ and $14\lambda_0$ from the source, whereas the far-field distance of the source array corresponds to $8\lambda_0$ for a 19-element CCA. The deflector element sizes are varied from $\frac{\lambda_0}{2}$ to $\frac{3\lambda_0}{2}$ to see their effect on the radiation patterns. The scan amplification factor $K_{\text{scan}} = 1.5$ is chosen to define the fixed phase-shifts of the deflector elements for constant gain over a wide angular coverage. The source antenna array is scanned to the maximum scan angle of $40^0$ and the operating frequency of 60 GHz is considered.

The performance of HTID is compared with a spherical deflector in Figure 10a. It shows that full hemispherical scan coverage can be achieved with spherical and HTID configurations, while satisfying the minimum directivity requirement of almost 10 dBi, with maximum source scan angle of almost 40°. The effect of increasing the dimension of deflector element to $\frac{3\lambda_0}{2}$ or more can be seen in Fig. 10b. The results show the better performance of a spherical deflector than the HTID with the realizable deflector element size of $\frac{\lambda_0}{2}$, in terms of radiation patterns. The dimension of larger than $\frac{\lambda_0}{2}$ doesn’t effectively populate the required facet boundary of HTID with deflector elements (Fig. 9). When multiple under-populated facets are combined for the HTID geometry, ripples are seen in the pattern due to less number of available phase states (missing deflector elements) leading to discontinuities. The effective population of facets with deflector elements can be carried out by increasing the radius of the circumscribed sphere. However, it will result in an increased phase-deviation leading to high-frequency ripples. Figure 11 shows the patterns for HTID having an average radius of $14\lambda_0$ with element sizes of $\frac{\lambda_0}{2}$ and $\frac{3\lambda_0}{2}$. The ripples can be seen in both the cases, although the fluctuation is less pronounced with element size of $\frac{\lambda_0}{2}$, in comparison with $\frac{3\lambda_0}{2}$ size. The results for 3D deflector geometries are consistent with the remarks made in Section III-B(3) for the heuristic approach of phase determination.

As determined from the simulation results of the analytical modelling, the directivity versus scan coverage envelopes of the source antenna array with and without spherical deflector are plotted in Fig. 12. Although the single planar antenna array has a high broadside directivity, however, once scan losses are considered, the spherical deflector configuration outperforms it for scan angles greater than almost 60°.

The variations of the radiation pattern of the 3D spherical passive electromagnetic deflector for the frequency range of 57 GHz to 64 GHz are also investigated. The analogue modulus $2\pi$ phase-shifters for the deflector elements instead of true-time delay behaviour are considered as they represent a more realistic and practical implementation of the deflector elements. The phase-shift distribution is depicted in Fig. 13a. The directivity patterns of a spherical deflector having modulus $2\pi$ phase-shifters for the deflector elements are shown in Fig. 13b. The decreased and increased values of directivity have been observed at 57 GHz and 64 GHz, respectively which is explained by the directivity values of the planar source antenna array at the respective frequencies. The directivity patterns for a spherical deflector show acceptable variation in angular coverage for the whole 60 GHz frequency band with the observation of some ripples for 57 GHz and 64 GHz frequencies.

VI. 60 GHZ DEFLECTOR ELEMENT

The previous sections provided the understanding and behaviour of the deflector configurations. The focus was on achieving wide angular coverage patterns using the appropriate phase-shift distribution on the deflector surface. However, no losses have been accounted for. An important aspect of the deflector configuration is the investigation of associated losses and design schemes for the realization of a deflector element with phase-shift range from $0^0$ to $360^0$ and minimal losses over the whole band of operation (57-64 GHz). The aforementioned defines the scope of this section with an emphasis
on the design of planar deflector elements having dimensions of close to $\lambda$. It is also desirable to avoid any active elements that can increase both the cost and the complexity of the deflector design. Moreover, the focus is on ‘low-profile’ deflector element design, realizable in printed circuit board (PCB) technology. The single/multilayer PCB technology facilitates simultaneous fabrication of many elements, unlike [21], where each element requires independent fabrication and assembling, leading to expensive and bulky structures. The phase-shifting property can be implemented using variable transmission-line lengths, element rotation or multiple resonant elements. The polarization schemes, for instance, linear-polarization (LP) or circular-polarization (CP), can also be incorporated in the deflector element design, either with the same or the different polarization schemes on the reception and the transmission sides. However, the use of CP on the transmission side can significantly increase the robustness of a wireless communication link, in comparison to LP, that may require accurate polarization alignment, especially for portable applications.

The focus of this section is on planar deflector elements, however, the presented schemes can be applied to conformal deflector elements as well. Table II broadly categorizes the reviewed literature in three design schemes, based on the realization of the phase-shifting property of the deflector element. The schemes are illustrated in Fig. 14. The specific implementation of each of the three design schemes, selected for 60 GHz frequency band are tabulated in Table III and shown in Fig. 15.

For fair comparison, the design and evaluation of the 60 GHz deflector elements are done using the Floquet analysis in commercially-available 3D EM tool, CST MWS. The frequency-domain solver has been used with ‘unit-cell’ boundary conditions to simulate an infinite array environment in CST MWS software having the provision of arbitrary incident angles for the plane-wave. The Floquet ports (1 and 2) are used which are similar to the waveguide ports in the sense that they absorb the radiation from an element as Floquet modes (which are plane-waves). This allows to obtain results similar to S-parameters, such as power transmission from the excited mode to the radiated mode(s). As an example, $S_{21}^{F,F}$ relates Fig. 9. Rectangular grid arrangement of deflector elements inside the polygonal facet boundary in LCS for deflector element sizes of a) $\frac{\lambda}{8}$, b) $\frac{\lambda}{6}$ (the black-dots represent selected deflector elements within the facet).

Fig. 10. Comparison of directivity (dBi) patterns ($\phi = 90^\circ$-plane, $f = 60$ GHz) for source-deflector geometry when a 19-element CCA source antenna array is scanned to $0^\circ$ [blue], $10^\circ$ [red], $20^\circ$ [black], $30^\circ$ [magenta] & $40^\circ$ [brown] a) source patterns [dotted], source-HTID configuration [solid] ($K_{scan} = 1.5$, $R = 7\lambda_0$ & $d = 0.16\lambda_0$) and spherical deflector [dashed] ($K_{scan} = 1.5$, $R = 7\lambda_0$ & $d \approx 0.16\lambda_0$) b) source-HTID configuration ($K_{scan} = 1.5$, $R = 7\lambda_0$) with $d = 0.5\lambda_0$ [solid] and spherical deflector [dashed] ($K_{scan} = 1.5$, $R = 7\lambda_0$ & $d \approx 0.5\lambda_0$).

Fig. 11. Comparison of directivity (dBi) patterns ($\phi = 90^\circ$-plane, $f = 60$ GHz) for source-HTID geometry when a 19-element CCA source antenna array is scanned to $0^\circ$ [blue], $10^\circ$ [red], $20^\circ$ [black], $30^\circ$ [magenta] & $40^\circ$ [brown] for $K_{scan} = 1.5$, $R = 14\lambda_0$ & $d = 0.16\lambda_0$ [dashed], $0.5\lambda_0$ [solid].
transmission between Floquet modes \( \tilde{F} \) and \( \hat{F} \) at ports 2 and 1, respectively, when port 1 is excited with mode \( \hat{F} \) (modes \( \tilde{F} \) and \( \hat{F} \) can be linearly-polarized or circularly-polarized). The CST simulation set-up is used to simulate the transmission magnitude and phase of all the three deflector elements.

**TABLE II**

<table>
<thead>
<tr>
<th>Scheme No.</th>
<th>Design schemes</th>
<th>Publication</th>
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<tbody>
<tr>
<td>I</td>
<td>Transmission-line</td>
<td>[15], [22], [23]</td>
</tr>
<tr>
<td>II</td>
<td>Element rotation</td>
<td>[24], [14]</td>
</tr>
<tr>
<td>III</td>
<td>Multiple resonant elements</td>
<td>[25], [26]</td>
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**Fig. 12.** Directivity versus scan coverage envelopes of the source antenna array without (□) and with the spherical deflector (○) for \( K_{scan} = 1.5 \). 

**A. I - Transmission-line based deflector element**

In transmission-line based LP deflector element design, the power from the reception side is coupled to the transmission side through a pair of transmission-lines and the second pair of slots, resulting in the orthogonal polarization of the emitted wave as compared to the incident wave. The transmission-lines of variable lengths have been used to achieve the phase-shift. The design details of the deflecting elements are mentioned in [15]. The three different deflector elements having the dimension of \( 0.6\lambda_0 \times 0.64\lambda_0 \) (\( \lambda_0 \) is the free-space wavelength at 60 GHz) realize a relative phase-shift of -120°, 0°, and +120° for the 57-63 GHz frequency band among themselves.

**TABLE III**

<table>
<thead>
<tr>
<th>Design scheme</th>
<th>Selected concept</th>
<th>Polarization</th>
</tr>
</thead>
<tbody>
<tr>
<td>II - Element rotation</td>
<td>[24]</td>
<td>RHCP</td>
</tr>
<tr>
<td>III - Multiple resonant elements</td>
<td>[25]</td>
<td>LP</td>
</tr>
</tbody>
</table>

**Fig. 13.** Spherical deflector (\( K_{scan} = 1.5, R = 9\lambda_0 \) & \( d \approx 0.5\lambda_0 \)) with modulus 2\( \pi \) phase-shifters for the deflector elements (a) Phase-shift (deg.), (b) Directivity (dBi) patterns for source-spherical deflector configuration, when a 19-element CCA source antenna array is scanned to 0°, 10°, 20°, 30°, 40° & 50° at 57 GHz [dashed], 60 GHz [solid] and 64 GHz [dotted].

**Fig. 14.** Realization of the phase-shifting property of a deflector element using (a) transmission-line, (b) element rotation and (c) multiple resonant elements design schemes.
The coupling of the pair of patches on either side of the ground comprises a crossed-shaped aperture in the ground plane for directing circular polarization with element rotation. The element provides phase-shift and a change in handedness of the incident circularly-polarized deflector element is based on [27], [28] and simulated for the 60 GHz frequency band. The presented B. II - Rotation based deflector element achieves the phase-shifting property. The principle is to design one element and rotate it to realize different phase-shifts by only varying transmission-line lengths because it affects the transmission properties due to coupling issues of transmission-line layer and slots layer. Moreover, it asks for a separate optimized design for each deflector element, having a specific phase-shift. Due to the presence of transmission lines, it is very difficult to restrict the deflector element’s dimension to 0.5λ₀, which is required in order to avoid onset of grating lobes that further degrades and limits the performance of the deflector element.

B. II - Rotation based deflector element

The circularly-polarized deflector element is designed and simulated for the 60 GHz frequency band. The presented 60 GHz CP deflector element is based on [27], [28] and provides phase-shift and a change in handedness of the incident circular polarization with element rotation. The element comprises a crossed-shaped aperture in the ground plane for coupling the pair of patches on either side of the ground plane, resulting in five metal layers. The use of two patches on either side increases the bandwidth of operation and the deflector element is matched to the incident plane wave by making the size of patches close to half-wavelength resonance and dimensioning the cross-shaped aperture. The deflector element’s dimensions does not exceed 0.5λ₀ x 0.5λ₀.

Figure 17a shows the variation of |S₂₁| parameter with angle of incidence for different azimuth angles 0° (solid), 45° (dashed), 90° (dashed-dotted) and 135° (dotted) (a) magnitude (dB), (b) phase (deg.). The above simulation results conclude that the proposed circularly-polarized deflector element operates successfully, in terms of realization of phase-shift range [0°,360°] and good transmission properties (> -1 dB) for the desired frequency band (57-64 GHz).

The 60 GHz deflector element, based on scheme-III (multiple resonant elements) is also designed, using the dual-resonant double square rings deflector element (DRDSR) [25]. The simulation results are highlighted in [17].

A comparison of all the three mentioned deflector element design schemes based on the simulation results for the 60 GHz frequency band is tabulated in Table IV.

VII. CONCLUSIONS

The pros and cons of the deflector configuration for wide angular coverage are clearly identified using the developed...
computationally-efficient analytical tool together with ‘small-scale’ 3D EM simulations. The said approach for the deflector analysis is seen as very useful in reducing the design space considerably for the computationally-very-intensive full-scale 3D EM simulation of the deflector configuration.

The deflector extends the scan-range of a steerable planar antenna array (source) by introducing phase-shifts on the deflecting surface, with the antenna gain high enough (> 10 dBi) to satisfy the minimum gain requirement over the whole coverage region. The results show that for the source’s scan angle of almost 40°, a hemispherical coverage can be achieved by the deflector, at the cost of average 3 dB gain loss and with almost a constant-gain envelope. The constant-gain envelope is required for the utilization of maximum allowed EIRP over the whole scan range. The minimum gain requirement over the hemispherical coverage can be fulfilled by adjusting (increasing/decreasing) the number of elements for the planar source antenna array of the combined source-deflector configuration. The analysis of the spherical/multifaceted deflector configurations shows the better performance of a spherical deflector than the multifaceted deflector counterpart with the realizable deflector element size of $\frac{\lambda}{4}$, in terms of radiation patterns.

A 60 GHz circularly-polarized deflector element has been designed which provides phase-shift with element rotation upon transmission of circular polarization. The simulation results show that the proposed circularly-polarized element operates successfully, in terms of realization of phase-shift range $[0^\circ, 360^\circ]$ and good transmission properties (> -1 dB) for the desired frequency band (57-64 GHz). The comparison of the three deflector element design schemes shows that the 60 GHz circularly-polarized deflector element based on element rotation scheme achieves the desired performance. Most importantly, with the single deflector element design, the required phase-shift range can be achieved by rotation, unlike the other design schemes.

The presented work provides a basic understanding and paves the way for the design of a full-fledged deflector ‘demonstrator’ with the proposed CP deflector elements and steerable circularly-polarized source array [29].

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